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NOISE IMMUNITY IN THE INTEGRAL RECEPTION OF SIGNALS ON A BACKGROUND OF SHORT SINUSOIDAL AND PULSED NOISE

N. L. Teplov

A comparison of the noise immunities of the methods of integral and ordinary narrowband reception of signals is given for the case where short-duration sinusoidal and pulsed noise occur. The effect of irregular changes in signal amplitude is taken into account. The prospects of integral methods of pulsed signal reception are evaluated.

NOISE IMMUNITY IN INTEGRAL AND NARROWBAND METHODS OF RECEPTION OF SIGNALS WITH PULSED NOISE

Narrowband Signal Reception. Pulsed noise at the output of any resonating system is a series of nonperiodic single pulses excited by short duration stray EMF's sufficiently dispersed in time. Alternatively, pulsed noise can be defined as the reaction of a filter to impulsive-type excitation by means of short-duration single EMF's. These EMF's can have the form of aperiodic or oscillating pulses; the disturbance time is less and the repetition rate is greater than the duration of nonstationary processes in the filter. In particular it follows from the above that pulse-type noises at the output of a filter possess a wider spectrum than its transmission band. In its general form pulsed noise at the output of a filter can be described by the equation

$$C(t) = c(t)\cos(\omega_0 t + \varphi_0), \tag{1}$$

where ω_0 is the angular tuning frequency of the filter; φ_0 is determined by the instant of appearance of the noise; c(t) is the amplitude (envelope) of the oscillations, the changes of which with time determine the form of the pulse interference at the output of the filter. Function c(t) will be determined by the frequency characteristics of the system itself in the case where the spectral density of each of the interference pulses is practically constant at the output of the resonating system and is within the limits of the transmission band of the latter.

Let us express the function c(t) for the two extreme cases [for example 4]:

1) for the interference at the output of a single tank circuit

$$c(t)_{\text{s.t.c.}} = 2S_0\Delta f_{\text{eff}} e^{-2\Delta f_{\text{eff}} t} \text{ for } 0 \leqslant t \leqslant \infty; \tag{2}$$

2) for the interference at the interference at the output of an ideal bandpass filter (with a rectangular resonance characteristic)

$$c(t)_{\text{ideal filter}} = 2S_0 \Delta f_{\text{eff}} \frac{\sin \pi \Delta f_{\text{eff}} t}{\pi \Delta f_{\text{eff}} t} \quad \text{for} \quad -\infty \leqslant t \leqslant \infty. \tag{3}$$

Here S_0 is the modulus of the spectral density of the pulse interference at the filter output which is taken as a constant within the limits of the passband of the filter; Δf_{eff} is the effective frequency band of the filter (for the case of an ideal resonating system the effective frequency band is identically equal to the transmission band Δf)[1].

The form of pulse interferences at the output of an ideal filter and a single tank circuit is shown in Figures 1a and b respectively.

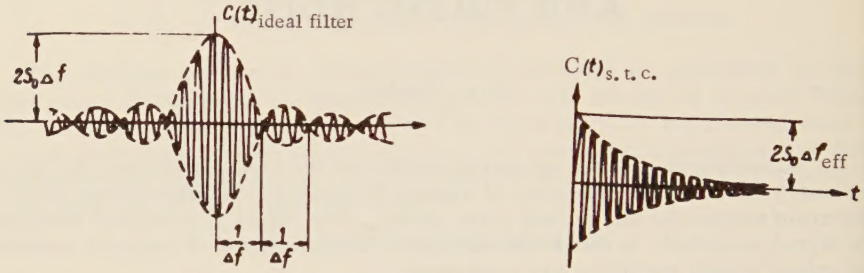

Figure 1

Pulse noise at the output of any resonating system has a maximum amplitude

$$c_{\text{out max}} = c(t)_{t=0} = 2S_0 \Delta f_{\text{eff}} \tag{4}$$

which is equal to twice the product of the spectral density of noise and the effective frequency bandwidth.

Let us determine the relation between the maximum amplitudes of the single pulse signal and the noise at the output of the narrowband filter.

For the single tank circuit this ratio is equal to

$$\left(\frac{a_{\text{out max}}}{c_{\text{out max}}}\right)_{\text{s.t.c.}} = \frac{a(1 - e^{-2\Delta f_{\text{eff}} \tau})}{2S_0 \Delta f_{\text{eff}}}. \tag{5}$$

In (5) the duration of the signal pulse is indicated by τ.

It can be easily seen that the ratio (5) increases with a decrease in the effective frequency bandwidth of the circuit. In the case of an unlimited narrow frequency band of the circuit

$$\lim_{\Delta f_{\text{eff}} \to 0} \frac{a_{\text{out max}}}{c_{\text{out max}}} = \frac{a}{2S_0} 2\tau = \frac{a\tau}{S_0}. \tag{6}$$

However, an excessive narrowing-down of the frequency bandwidth during ordinary narrowband reception is contradictory to the requirements of maximum suppression of other forms of noise, particularly fluctuating and "nonattenuated" sinusoidal noise for which the optimum frequency bandwidths are indicated in [2]. The reception of signals in the form of an

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[1]The coefficient of transmission of filters at the resonant frequency has been arbitrarily set equal to unity.

uninterrupted sequence of pulses is entirely impossible in the case of excessively narrow transmission bands, since, when the pulses mutually overlap, the probability of their detection falls off sharply.

For $\Delta f_{\text{eff opt}} = 0.65/\tau$ (for single pulse signals)

$$\left(\frac{a_{\text{out max}}}{c_{\text{out max}}}\right)_{\text{opt s. t. c.}} = \frac{a}{S_0}\tau\frac{(1-e^{-1.3})}{2\cdot 0.65} = 0.56\,\frac{a\tau}{S_0}. \tag{7}$$

For $\Delta f'_{\text{eff opt}} = 1.1/\tau$ (during the reception of an uninterrupted sequence of pulses)

$$\left(\frac{a_{\text{out max}}}{c_{\text{out max}}}\right)_{\text{opt}'\text{ s. t. c.}} = 0.36\,\frac{a\tau}{S_0}. \tag{8}$$

Analogous results are obtained for an ideal bandpass filter for which

$$\left(\frac{a_{\text{out max}}}{c_{\text{out max}}}\right)_{\text{ideal filter}} = \frac{\frac{2}{\pi}\,a\,\text{SI}(\pi\Delta f\tau)}{2S_0\Delta f}. \tag{9}$$

For $$\Delta f_{\text{opt}} = \frac{1.37}{\tau}$$

$$\left(\frac{a_{\text{out max}}}{c_{\text{out max}}}\right)_{\text{opt ideal filter}} = \frac{\frac{2}{\pi}\,a\,\text{SI}(1.37\pi)}{2S_0\frac{1.37}{\tau}} \approx 0.4\,\frac{a\tau}{S_0}. \tag{10}$$

And correspondingly

$$\left(\frac{a_{\text{out max}}}{c_{\text{out max}}}\right)_{\text{opt}'\text{ ideal filter}} \approx 0{,}25\,\frac{a\tau}{S_0}. \tag{11}$$

In the general case we may assume

$$\left(\frac{a_{\text{out max}}}{c_{\text{out max}}}\right)_{\text{opt}} \approx 0.5\,\frac{a\tau}{S_0}; \tag{12}$$

$$\left(\frac{a_{\text{out max}}}{c_{\text{out max}}}\right)_{\text{opt}'} \approx 0.3\,\frac{a\tau}{S_0}. \tag{13}$$

Integral Reception of Signals. During integration the integrator is a resonating system (filter) with a very narrow static transmission band. Self-oscillations are excited in this filter by the action of a short duration interference pulse. The maximum amplitude of these oscillations (independent of the type of integrating circuit) is determined by formula (4), where Δf_{eff} is the effective static band of the integrator filter. In the case of ideal integration the static frequency bandwidth of the integrator filter approaches zero ($\Delta f_{\text{eff}} \to 0$). From this one can conclude that in the case of integral reception the maximum ratio of the signal to the pulse interference is attained; this is determined by formula (6),

$$\left(\frac{a}{c}\right)_{\text{int}} = \frac{a\tau}{S_0}. \tag{14}$$

The result of (14) can be obtained in another manner.

In the integral reception scheme, a channel with a relatively wide frequency bandwidth precedes the integrator. The interference pulses, the form of which is determined by the frequency characteristics of the channel, are excited at the output of the latter. These interference pulses are then subjected to integration in the integrator. It can be easily shown that the result of the integration does not depend on the form of the interference pulse and can be expressed by relation (14) in any case.

Actually the amplitude of the signal at the output of the integrator is equal to

$$a_{\text{int}} = a\tau. \tag{15}$$

The integral value of the interference pulse amplitude is equal to:

a) in the case of a wideband channel in the form of a single tank [on the basis of formula (2)]

$$c_{\text{int}} = \int_0^\infty 2S_0 \Delta f_{\text{eff}} e^{-2\Delta f_{\text{eff}} t} dt; \tag{16}$$

b) in the case of a wideband channel in the form of an ideal band filter [on the basis of formula (3)]

$$c_{\text{int}} = \int_{-\infty}^{\infty} 2S_0 \Delta f_{\text{eff}} \frac{\sin \pi \Delta f t}{\pi \Delta f t} dt. \tag{17}$$

In formulas (16) and (17) the integration within the limits of the duration of signal τ is replaced by integration over an infinite time on the assumption that the attenuation time for the interference pulses is considerably shorter than the duration of signal τ (which corresponds to a relatively wide band of channel frequencies for the receiver at the output of the integrator compared to $1/\tau$).

The computation of the integrals in formulas (16) and (17) yields

$$c_{\text{int}} = S_0. \tag{18}$$

The ratio of the signal (15) to the interference (18) equals

$$\left(\frac{a}{c}\right)_{\text{int}} = \frac{a\tau}{S_0}, \tag{19}$$

which is identical to formula (14). In this manner the ratio of the signal amplitude to the pulse interference amplitude at the output of the integrator is equal to the ratio of the product of the signal amplitude and its duration to the spectral density of the noise.

Let us compare the ratios of signal-to-pulse interference during integral reception and ordinary narrowband reception with optimum frequency bandwidths. Comparing formula (14) with formulas (12) and (13), we obtain:

a) during the reception of single pulses

$$\frac{\left(\frac{a}{c}\right)_{\text{int}}}{\left(\frac{a}{c}\right)_{\text{opt}}} = 2; \tag{20}$$

b) during the reception of an uninterrupted sequence of pulses

$$\frac{\left(\frac{a}{c}\right)_{\text{int}}}{\left(\frac{a}{c}\right)_{\text{opt}'}} \approx 3. \tag{21}$$

Consequently the gain in interference immunity attained by the integral method of reception during the action of pulse interferences, as compared to the ordinary narrowband reception with optimum frequency bandwidths, turns out to be equal to two and three with respect to voltage (four and ten with respect to power) during the reception of single and uninterrupted sequency of pulses respectively.

NOISE IMMUNITY IN INTEGRAL AND NARROWBAND RECEPTION WITH SHORT-DURATION SINUSOIDAL NOISE

Integral Reception of Signals. Sinusoidal noise is equally probable at any frequency. We shall assume that the average statistical value of the amplitude of the noise at any frequency will be the same and equal to e'_n.

Then the average ensemble power of short-duration sinusoidal noises, each with a duration of τ_n, at the output of the integrator [2] will be equal (across a resistor of 1 ohm) to

$$\overline{e_n^2} = \int_{-\infty}^{\infty} (e'_n \tau_n)^2 \frac{\sin^2 \pi \Delta f \tau_n}{(\pi \Delta f \tau_n)^2} d(\Delta f) = e_n'^2 \tau_n = 2\xi_0^2 \tau_n, \tag{22}$$

where ξ_0^2 is the square of the effective value of the sinusoidal noise assumed to be equal at any frequency.

Consequently the ratio of signal to short-duration sinusoidally ideal noise during integral reception is obtained as equal to

$$h_*^2 = \frac{a^2_{int}}{\overline{e_n^2}} = \frac{a^2\tau}{e_n'^2} \frac{\tau}{\tau_n} = \frac{a^2\tau}{2\xi_0^2} \frac{\tau}{\tau_n}. \tag{23}$$

When $\tau_n = \tau$

$$h_*^2 = \frac{a^2\tau}{2\xi_0^2}, \tag{24}$$

which coincides identically with the ratios of signal and "nonattenuated" sinusoidal noise at the output of the integrator [2].

Narrowband Reception of Signals. Let us determine the average ensemble power of short-duration sinusoidal noises at the output of a single tank circuit.

The amplitude of the noise, at frequency f, at the end of its duration $t = \tau_n$ is equal to [3] [2]

$$e_{n\,out}(\tau_n) = \frac{e'_n}{\sqrt{1 + \left(\pi \frac{\Delta f}{\Delta f_{eff}}\right)^2}} \sqrt{1 - 2e^{-2\Delta f_{eff}\tau_n} \cos 2\pi \Delta f \tau_n + e^{-4\Delta f_{eff}\tau_n}}. \tag{25}$$

Here $\Delta f = |f - f_0|$ is the absolute value of the detuning of interference f, relative to the resonant frequency of the circuit tuned to the signal frequency f_0; Δf_{eff} is the effective frequency bandwidth of the circuit.

Then the average ensemble power of the short-duration sinusoidal noise at the output of the circuit is determined from the expression

$$\overline{e_{n\,out}^2} = e_n'^2 \, 2 \int_0^{\infty} \frac{1}{1 + \left(\pi \frac{\Delta f}{\Delta f_{eff}}\right)^2} \times \times (1 - 2e^{-2\tau_n \Delta f_{eff}} 2\cos \pi \tau_n \Delta f + e^{-4\tau_n \Delta f_{eff}}) d(\Delta f). \tag{26}$$

After computing the integral we obtain

$$\overline{e_{n\,out}^2} = e_n'^2 \Delta f_{eff} (1 - e^{-4\Delta f_{eff}\tau_n}). \tag{27}$$

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[2] The coefficient of transmission of the circuit at the resonating frequency is conditionally assumed equal to unity.

The amplitude of the signal at the output of the circuit is

$$a_{out} = a\left(1 - e^{-2\Delta f_{eff}\tau}\right). \tag{28}$$

In this manner the signal-to-noise ratio at the output of a narrowband resonating circuit turns out to be equal to

$$\left(\frac{a^2}{\bar{e}_n^2}\right)_{out} = \frac{a^2\left(1 - e^{-2\Delta f_{eff}\tau}\right)^2}{e_n'^2\,\Delta f_{eff}\left(1 - e^{-4\Delta f_{eff}\tau_n}\right)}. \tag{29}$$

In the case of an unlimited narrowing of the transmission band of the circuit we obtain

$$\lim_{\Delta f_{eff}\to 0} \frac{a_{out}^2}{\bar{e}_{n\,out}^2} = \frac{a^2\tau}{e_n'^2}\,\frac{\tau}{\tau_n} = \frac{a^2\tau}{2\xi_0^2}\,\frac{\tau}{\tau_n}, \tag{30}$$

i. e., a result identical to the result obtained for integral reception.

It was pointed out above that it is impossible to attain excessively narrow transmission bands for a sequence of pulses without lengthy intervals (particularly in radio telegraphy).

Let

$$\Delta f_{eff\,opt} = \frac{0.65}{\tau}$$

(optimum frequency bandwidth of the circuit for fluctuating and "nonattenuated" sinusoidal noise during the reception of single pulses).

Then

$$\left(\frac{a_{out}^2}{\bar{e}_{n\,out}^2}\right)_{opt} = \frac{a^2\left(1 - e^{-1.3}\right)^2}{e_n'^2 \cdot \frac{0.65}{\tau}\left(1 - e^{-2.6\tau_n/\tau}\right)} = \frac{a^2\tau}{e_n'^2}\,\frac{0.82}{1 - e^{-2.6\tau_n/\tau}}. \tag{31}$$

Let us compute the relative signal-to-noise ratio for the case of integral reception (23) and narrowband reception (31):

$$A = \frac{(23)}{(31)} = \frac{\tau}{\tau_n}\,\frac{1 - e^{-2.6\tau_n/\tau}}{0.82}, \tag{32}$$

$$A = 1.22\left(1 - e^{-2.6p}\right)\frac{1}{p}, \tag{33}$$

where $p = \tau_n/\tau$ the ratio of the duration of the interference to the signal.

When $p = 1$

$$A = 1.22 \cdot 0.93 = 1.13. \tag{34}$$

When $p \to 0$

$$\lim_{p\to 0} A = 1.22 \lim_{p\to 0} \frac{1}{p}\left(1 - e^{-2.6p}\right) = 1.22 \cdot 2.6 = 3.2. \tag{35}$$

In the reception of pulses, which follow without intervals, the optimum frequency bandwidth, as was shown in [2], is equal to

$$\Delta f'_{eff\,opt} = \frac{1.1}{\tau}.$$

The rise in amplitude of the signal, meanwhile, is determined by the formula

$$\Delta a = a\left(1 - e^{-2\Delta f'_{eff\,opt}\tau}\right) - a\left(1 - e^{-2\Delta f'_{eff\,opt}\tau}\right)e^{-2\Delta f'_{eff\,opt}\tau} =$$

$$= a\left(1 - e^{-2\Delta f'_{eff\,opt}\tau}\right)^2. \tag{36}$$

The relative signal-to-noise ratio of integral (23) and narrowband (29) reception turns out to be equal, in the considered case, to

$$A' = \frac{(23)}{(29)_{\Delta f_{\text{eff}} = \Delta f'_{\text{eff opt}} = 1.1/\tau}} = \frac{1}{p} \frac{1{,}1\,(1 - e^{-4{,}4p})}{(1 - e^{-2{,}2})^4} = 1.75 \frac{1}{p} (1 - e^{-4{,}4p}); \quad (37)$$

here, instead of a, the amplitude of the signal, Δa, the value of its relative rise as determined by formula (36), was substituted into formula (29).

When p = 1

$$A' = 1.73. \quad (38)$$

When $p \to 0$

$$\lim_{p \to 0} A' = 1.75 \cdot 4.4 = 7.7. \quad (39)$$

It can be easily established that when $p \to 0$ the ratio of signal to sinusoidal noise during integral and narrowband methods of reception approaches asymptotically to the ratio of signal-to-pulse interference using the same methods of reception. Thus, taking the ratio of the formulas (14) and (7) (when $\Delta f_{\text{eff opt}} = 0.65/\tau$) and (14) and (8) (when $\Delta f'_{\text{eff opt}} = 1.1/\tau$) and raising these ratios to the second power, we obtain results which are identical to (35) and (39). Let us remember that the latter identities are directly applicable to the case where a single tank circuit is taken as an example of a narrowband resonating system. Analogous identities can also be established for the case of a narrowband channel with an ideal resonating characteristic.

On the other hand, when p = 1 the ratios of signals to sinusoidal noise during integral and narrowband methods of reception [formulas (34) and (38)] turn out to be practically equal to corresponding ratios during the action of fluctuating noise [2].

On the basis of the above statements one can finally write

$$\left(\frac{h_*^2}{h^2_{\text{opt}}}\right)_{\text{sinus.}} \approx [1.2 - 4]_{p=0-1}; \quad (40)$$

$$\left(\frac{h_*^2}{h^2_{\text{opt}}}\right)_{\text{sinus.}} \approx [2 - 10]_{p=0-1}. \quad (41)$$

Consequently, depending upon the ratio of the duration of the sinusoidal noise and the signal, the gain during integral reception is evaluated by the increase of the ratio of the powers of signal to noise of 1.2 to 4 times during the reception of single signal pulses and of 2 to 10 times during the reception of an uninterrupted sequence of pulses.

EFFECT OF IRREGULAR CHANGES OF THE PULSE SIGNAL AMPLITUDE

During ordinary narrowband reception of an uninterrupted sequence of pulses the accuracy of detection of the latter is decreased due to their mutual overlap. The accuracy of detection is further decreased during reception of signal pulses with a variable amplitude (for example, those subjected to attenuations). Such irregular changes in the amplitude of the pulse signals have the greatest effect when the amplitudes of adjacent pulses vary independently of one another. Such irregular changes, for example, occur during selective attenuations of the signal which take place when separate signal pulses are transmitted on different frequencies.

Let it be assumed that the reception of an uninterrupted sequence of

pulses whose amplitudes vary independently is occurring. As before, we will assume that the registering of each of the received signals takes place at the moment when its amplitude is at its maximum.

During ordinary narrowband reception at the instant of registration remnant oscillations of the preceding signal exist along with the oscillations of the former. Relative value of the amplitude of these oscillations will be increased with a narrowing of the transmission band of a narrowband filter and by the relative decrease of the received pulse amplitude over the amplitude of the previous pulse.

For the sake of simplicity let us assume that a single tank circuit acts as a narrowband filter and that the ratio of the amplitudes of the previous and the received pulses is equal to (Figure 2)

$$k = \frac{a'}{a}.$$

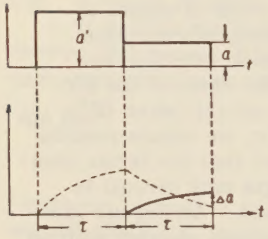

Figure 2

As before, let it be that the criterion for the evaluation of the interference immunity is the ratio of Δa, the rise in signal amplitude, to the average value of the amplitude of the interference. Then, using the methods [2], one can determine the optimum values of the effective frequency bandwidths taking into account the overlap of pulses and also their corresponding ratios of signal to noise for each fixed ratio of amplitudes of adjacent pulses k.

In this case (see Figure 2)

$$\Delta a_{max} = a\left(1 - e^{-2\Delta f''_{eff\,opt}\tau}\right) - ka\left(1 - e^{-2\Delta f''_{eff\,opt}\tau}\right)e^{-2\Delta f''_{eff\,opt}\tau}, \qquad (42)$$

where $\Delta f''_{eff\ opt}$ is the optimum frequency bandwidth taking into account the mutual overlap of pulses when the ratio of the amplitudes of the previous and the received pulse is equal to k.

The results of such computations are shown in Table 1.

The following designations were used in Table 1:

$$Q_1^2 = \frac{a^2\tau}{2\nu_0^2}, \qquad (43)$$

where ν_0^2 is the specific (related to unit bandwidth) intensity of the fluctuating noise;

$$Q_2^2 = \frac{a^2\tau}{2\xi_0^2}, \qquad (44)$$

where ξ_0^2 is the square of the effective value of the "nonattenuated" sinusoidal noise of one of the frequencies

$$Q_3^2 = \frac{a^2\tau}{2\xi_0^2}\frac{\tau}{\tau_n}, \qquad (45)$$

where ξ_0^2 is the square of the effective value of the short-duration sinusoidal noise,

$$Q_4^2 = \frac{(a\tau)^2}{2\varepsilon_0^2}, \qquad (46)$$

where

$$\varepsilon_0^2 = \frac{S_0^2}{2}.$$

Table 1

Form of reception	k	$\Delta f_{eff\ opt}$	Ratio of signal to noise h^2			
			fluctuating noise	sinusoidal "non-attenuated" noise ($p \gg 1$)	short-duration sinusoidal noise ($p \ll 1$)	pulse interferences
Integral reception	any	$\left(\frac{1}{\tau}\right)$	Q_1^2	Q_2^2	Q_3^2	Q_4^2
Narrowband reception of single pulses	0	$0.65\frac{1}{\tau}$	$0.82\,Q_1^2$	$0.82\,Q_2^2$	$0.313\,Q_3^2$	$0.313\,Q_4^2$
Narrowband reception of uninterrupted sequence of pulses with constant amplitude	1	$1.1\frac{1}{\tau}$	$0.58\,Q_1^2$	$0.58\,Q_2^2$	$0.13\,Q_3^2$	$0.13\,Q_4^2$
Narrowband reception of an uninterrupted sequence of pulses with unequal amplitude	2	$1.5\frac{1}{\tau}$	$0.46\,Q_1^2$	$0.46\,Q_2^2$	$0.0805\,Q_3^2$	$0.0805\,Q_4^2$
	4	$1.7\frac{1}{\tau}$	$0.42\,Q_1^2$	$0.42\,Q_2^2$	$0,058\,Q_3^2$	$0.058\,Q_4^2$
	10	$2.2\frac{1}{\tau}$	$0.32\,Q_1^2$	$0.32\,Q_2^2$	$0.04\,Q_3^2$	$0.04\,Q_4^2$

Table 2

Form of reception	k	$\Delta f_{eff\ opt}$	Ratio of ratios h_*^2/h_{opt}^2		
			fluctuating noise	sinusoidal noise ($p=0-\infty$)	pulse interferences
Integral reception	any	$\left(\frac{1}{\tau}\right)$	1	1	1
Narrowband reception of single pulses	0	$0.65\frac{1}{\tau}$	1.22	1.22–3,2	3.2
Narrowband reception of an uninterrupted sequence of pulses with constant amplitude	1	$1.1\frac{1}{\tau}$	1.73	1.73–7.7	7.7
Narrowband reception of an uninterrupted sequence of pulses with unequal amplitude	2	$1.5\frac{1}{\tau}$	2.15	2.15–12.4	12.4
	4	$1.7\frac{1}{\tau}$	2.41	2.41 17.1	17.1
	10	$2.2\frac{1}{\tau}$	3.18	3.18 - 25.0	25.0

In formulas (43)-(46), a and τ are the amplitude and duration of the signal pulse, respectively.

Table 2 shows the relative signal-to-noise ratios of integral and ordinary narrowband forms of reception. In this manner the relative gain obtained by integral reception of signals will turn out that much larger, the greater is the drop in amplitude between the previous and the received pulse.

Usually amplitude peaks are eliminated in pulse signal reception schemes by means of limiting of pulse amplitudes. However the introduction of a principally nonlinear element into a receiving channel in order to level off signal amplitudes leads to a decrease in receiver interference immunity for sinusoidal and fluctuating noise.

During integral reception mutual interference between pulses is eliminated by forced quenching of the oscillations at the end of each pulse. In this case the ratios between the amplitudes of adjacent pulses do not have any meaning in the sense of the discussion above. Then it becomes possible to design a linear radio frequency channel for a receiver in which (during integral reception) a predominance of the signal over the interference is attained, as per the first line of Table 1.

CONCLUSIONS

One can formulate the following general conclusions:

1. Integral reception simultaneously (congruently) satisfies the requirements for reducing various forms of noise. In ordinary narrowband reception, attempts to decrease the intensity of pulse and short-duration sinusoidal noise by narrowing the transmission band of the receiver contravene the requirements of reducing the effects of irregular changes of the signal amplitude. In the case of an excessive narrowing of the frequency bandwidth (below the optimum) in addition, the requirements for reducing the fluctuating and "nonattenuated" sinusoidal noise are violated.

2. The noise immunity of integral reception, compared to ordinary narrowband reception, turns out to be higher for pulse and short-duration noise. During the reception of single pulses of the signal, the noise immunity of integral reception, with respect to fluctuating and "nonattenuated" sinusoidal noise differs insignificantly from the noise immunity of the ordinary narrowband reception with optimum frequency bandwidths (with respect to the ratio of the powers of the signal and the noise by a factor of 1.2). However for this value of the bandwidths, the noise immunity of ordinary narrow band reception, with respect to short duration sinusoidal and pulse-type noise, turns out to be significantly lower. The noise immunity of ordinary narrowband reception, in comparison to the integral method of reception, is lowered in addition for all forms of noise during irregular changes in signal amplitude.

3. The noise immunity of integral reception for each of the types of noise is determined by the signal-to-noise ratios as indicated in Tables 1 and 2. On the basis of the conducted analysis it can be stated that the values of ratios for integral reception given in the tables are most probable during division of the signal and noise by methods of linear selection (into which, in particular, we may include the methods of filtration in connection with ordinary narrowband reception, and averaging during integral reception). On this basis the signal-to-noise ratios obtained during integration of the signal and noise which are recorded on the first line of Table 1 can be called the limiting values. In particular, the limiting ratios for fluctuating noise correspond to maximum noise immunity for

incoherent reception of pulse signals, and if the phase of the received signals is known, and it becomes possible to eliminate the incoherent component of the noise, it corresponds to the potential noise immunity[1].

Note: One should remember that for sinusoidal noise the values of the ratios indicated in Tables 1 and 2 for the method of integral reception are the total maximum, i.e., the maximum over the entire ensemble of sinusoidal noises which are equiprobable with respect to frequency and amplitude. In each nonindividual case (during the action of single sinusoidal noise) the signal-to-noise ratio in the case of integral reception is formed corresponding to the resonance curve of the integrator (Figure 6, in [2]), the characteristic property of which is the presence of zero minima during detuning $|\Delta f| = n\ 1/\tau$ and maxima during the detuning $|\Delta f| \approx (n + 1/2)\ 1/\tau$, where $n = 1, 2, 3, \ldots$ The most unfavorable ratio of the signal to sinusoidal noise will occur for detunings corresponding to the maxima of the resonance characteristics. This property of the resonance characteristic of the integrator must be taken into account in the formulation of the requirements for the selectivity of the preliminary channel of the receiver (at the output of the integrator).

In connection with this it should also be pointed out that the presence of zero minima in the resonance characteristic of the integrator ensures the possibility of a practically complete division of the signals distributed with respect to frequency over an interval divisible by $1/\tau$. This peculiarity is directly attained in several versions of radio telegraph communication apparatus.

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REGARDING THE EXPEDIENCY OF MULTIFREQUENCY COMMUNICATION APPLICATIONS

P. A. Konstantinov

The interference immunity of multifrequency communication systems with automatic reception is discussed, in which signals with several different values of frequency are employed. A procedure is given for determining the required frequency stability at which the application of multifrequency communication systems becomes expedient in actual practice.

Binary coding is widely used in radio communication systems today. In this case the coding combinations are composed of elementary signals with two different values. It is possible to build up communication systems employing codes with a higher base; this becomes possible when the code combinations are composed of elementary signals with the number of different values exceeding two.

The procedure for determining the required frequency stability at which the application of multifrequency communication systems becomes justified in actual practice is given below.

THE RELATIVE PROBABILITY OF DISTORTION OF THE COMMAND IN LIMITING CASES

The total number of possible combinations in a binary code and in a code with a base m are, respectively,

$$N_2 = 2^{n_2}, \tag{1}$$

$$N_m = m^{n_m}, \tag{2}$$

where n_2 and n_m are the valences of the indicated codes.

Assuming that the total number of possible combinations is identical for different code bases, we obtain from Equations (1) and (2)

$$n_m = \frac{n_2}{\log_2 m}. \tag{3}$$

It can be seen from this that the number of elements n_m in a code based on m is smaller by a factor $\log_2 m$ than the number of elements n_2 in a binary code.

The probability of distortion of the elementary signal during the use of codes with bases m and 2 will be designated by P_m and P_2, respectively. Then the probability of correct reception of the unit signal will be $(1 - P_m)$,

and the probability of correct reception of all of the n_m unit signals which make up the code combination i.e., the probability of a correct reception of the command, becomes

$$Q_{cm} = (1 - P_m)^{n_m} \tag{4}$$

The probability of distortion of the command is

$$P_{cm} = 1 - (1 - P_m)^{nm} = n_m P_m - \frac{n_m (n_m - 1)}{2!} P_m^2 + \dots$$

In real communication lines we may assume that $P_m \ll 1$; therefore, the terms containing powers of P_m higher than the first can be neglected. Then

$$P_{cm} = n_m P_m. \tag{5}$$

Assuming that m = 2 we obtain

$$P_{c2} = n_2 P_2. \tag{6}$$

From (5), (6) and (3) we have

$$\frac{P_{cm}}{P_{c2}} = \frac{1}{\log_2 m} \frac{P_m}{P_2}. \tag{7}$$

Expression (7) relates the ratio of the probabilities of distortion of the command in various code bases to the ratio of the probabilities of distortion of the elementary signals.

The value of probability of distortion of the elementary signal depends upon the type of keying and the duration of the signal. The latter should be considered for the case where the probabilities of distortion of the command are compared for cases with equal command duration. When the base of the code is raised, the number of elementary signals in the code combination is decreased in accordance with Equation (3). This means that in order to realize a given duration of the command during the transfer from binary code to a code with a base m, the duration of the elementary signal can be increased by a factor of $\log_2 m$, i.e.,

$$\tau_m = \tau_2 \log_2 m, \tag{8}$$

where τ_m and τ_2 are the durations of elementary signals for code bases m and 2, respectively. The lengthening of the elementary signal during the raising of the base of the code and a sufficiently high frequency stability will permit a narrowing of the passband of the receiver, as a result of which the probability of distortion P_m in formula (7) is decreased.

The probability of distortion of the elementary signal and the command depends upon the form of the signal and the method of reception. The material presented below refers to multifrequency communication systems in which the code combinations are made up of a series of emitted parcels while to each of these there may correspond the oscillations of one of the m frequencies. It is assumed that reception is conducted in accordance with the method of tests and that the receiver contains a filter frequency detector with a comparison device at its output (Figure 1). Thus, the parcel is registered in that channel in which the voltage is highest.

For the considered case, L. M. Fink has obtained the formulas for the probability of correct reception of the elementary signal:

$$Q_m = 1 + \sum_{i=1}^{m-1} (-1)^i \frac{(m-1)!}{(i+1)!\,(m-i-1)!} \exp\left[-\frac{z_m^2 i}{2(i+1)}\right], \tag{9}$$

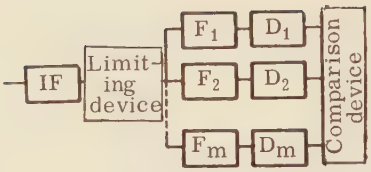

Figure 1

$$Q_2 = 1 - \frac{1}{2}\exp\left(-\frac{\alpha_2^2}{4}\right); \quad (10)$$

here $\alpha_m = A/\sigma_m$, $\alpha_2 = A/\sigma_2$ are the signal-to-noise ratios for code bases m and 2, respectively. On the basis of (9) and (10), and recalling that $P_m = 1 - Q_m$, expressions for the probability of distortion of the elementary signal may be written in the following manner:

$$P_m = (m-1)!\sum_{i=1}^{m-1}\frac{(-1)^{i+1}}{(i+1)!(m-i-1)!}\exp\left[-\frac{z_m^2 i}{2(i+1)}\right], \quad (11)$$

$$P_2 = \frac{1}{2}\exp\left(-\frac{z_2^2}{4}\right). \quad (12)$$

Substituting (11) and (12) into (7) we obtain

$$\frac{P_{cm}}{P_{c2}} = \frac{2(m-1)!}{\log_2 m}\exp\left(\frac{z_2^2}{4}\right)\sum_{i=1}^{m-1}\frac{(-1)^{i+1}}{(i+1)!(m-i-1)!}\exp\left[-\frac{z_m^2 i}{2(i+1)}\right]. \quad (13)$$

The relation between the quantities α_m and α_2, appearing in (13), may vary. At low frequency stability, when the transmission band of filter F_1, $F_2, \ldots, F_m$ depends only to a small degree upon the duration of the elementary signal, and is basically determined by the magnitude of instability, we may assume that $\alpha_m = \alpha_2$. Then from (13) we obtain

$$\frac{P_{cm}}{P_{c2}} = \frac{2(m-1)!}{\log_2 m}\sum_{i=1}^{m-1}\frac{(-1)^{i+1}}{(i+1)!(m-i+1)!}\exp\left[\frac{z_2^2(1-i)}{4(1+i)}\right]. \quad (14)$$

For high frequency stability and lengthening of the elementary signals in accordance with (8) by a factor of $\log_2 m$ will also lead to a similar narrowing of the transmission bands of the filters F_1, $F_2, \ldots, F_m$ and to an increase of the signal-to-noise ratio at the output of these filters. Therefore, in formula (13) we must assume that $\alpha_m = \alpha_2\sqrt{\log_2 m}$. In this case the relative probability of distortion of the command is equal to

$$\frac{P_{cm}}{P_{c2}} = \frac{2(m-1)!}{\log_2 m}\sum_{i=1}^{m-1}\frac{(-1)^{i+1}}{(i+1)!(m-i-1)!}\exp\left\{\frac{z_2^2[1+i(1-2\log_2 m)]}{4(i+1)}\right\}. \quad (15)$$

Formulas (14) and (15), as well as formula (7), hold for small values of P_m and P_2, i.e., at large values of α_2. When $\alpha_2 \gg 1$ we may neglect all but the first term, corresponding to i = 1; thus the indicated relations can be simplified to a considerable extent. In this case we obtain from (14) and (15), respectively:

$$\frac{P_{cm}}{P_{c2}} = \frac{m-1}{\log_2 m}, \quad (16)$$

$$\frac{P_{cm}}{P_{c2}} = \frac{m-1}{\log_2 m}\exp\left[\frac{z_2^2}{4}(1-\log_2 m)\right]. \quad (17)$$

The obtained formulas permit us to evaluate the relative probability of distortion of the command in the two extreme cases: for low and high

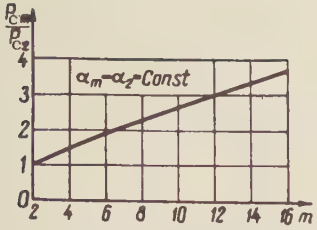

Figure 2

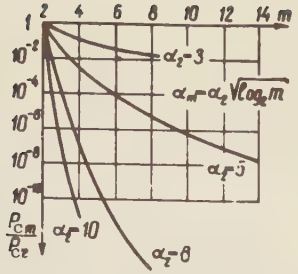

Figure 3

frequency stability. Figure 2 shows the dependence corresponding to formula (16) which applies to the low frequency stability case. It can be seen from the diagram that the raising of the base of the code leads to an insignificant increase in the probability of distortion of the command. Thus, in a transfer from binary to quaternary code, the probability of distortion of the command increases by a factor of 1.5 and during transfer to an octavalent code it increases by a factor of 2.3.

For high frequency stability, the probability of distortion of the command is governed by formula (17). The results of computations using this formula are shown in Figure 3; here it can be seen that the raising of the code base will lead to a sharp decrease in the probability of distortion of the command.

DEPENDENCE OF RELATIVE PROBABILITY OF DISTORTION OF THE COMMAND ON FREQUENCY STABILITY

Two extreme cases were discussed above: where the transmission band of the receiver (dividing filters) is determined entirely either by the frequency instability ($\alpha_m = \alpha_2 = \text{const}$), or by the duration of the elementary signal ($\alpha_m = \alpha_2 \sqrt{\log_2 m}$). In the first case the raising of the code base leads to a certain increase in the probability of distortion of the command, but in the second case it leads to a sharp decrease. In actual practice an intermediate case is of interest, where, when the code base is raised and the elementary signal is lengthened by a factor of $\log_2 m$, the frequency stability permits only a certain narrowing of the passband of the receiver which, in comparison with the passband in the binary code, is less than a factor of $\log_2 m$.

Let us assume that in raising the base of the code the duration of the elementary signal increases in accordance with the relation (8), i.e., $\tau_m = \tau_2 \log_2 m$ and the transmission band of the dividing filters is narrowed by a factor of q, i.e.,

$$\frac{\Delta f_{F2}}{\Delta f_{Fm}} = q.$$

In this case the quantity α_m increases by a factor of $\sqrt{q}$, i.e., $\alpha_m = \alpha_2 \sqrt{q}$ due to a decrease in the noise level.

Considering that

$$\frac{\alpha_2^2}{4} - \frac{\alpha_m^2 i}{2(i+1)} = \frac{\alpha_2^2}{4}\left(1 - \frac{2qi}{i+1}\right) = \frac{\alpha_2^2 [1 + i(1-2q)]}{4(i+1)},$$

we transform formula (13):

$$\frac{P_{Cm}}{P_{C2}} = \frac{2(m-1)!}{\log_2 m} \sum_{i=1}^{m-1} \frac{(-1)^{i+1}}{(i+1)!(m-i-1)!} \exp\left\{\frac{a_2^2[1+i(1-2q)]}{4(i+1)}\right\}. \quad (18)$$

Considering only the first term, we obtain the simplified formula

$$\frac{P_{Cm}}{P_{C2}} = \frac{m-1}{\log_2 m} \exp\left[\frac{a_2^2}{4}(1-q)\right]. \quad (19)$$

Equations (16) and (17) given previously are particular cases of Equation (19); (16) is obtained when q = 1 (the case of low frequency stability); (17) holds when $q = \log_2 m$ (the high frequency stability case).

In order to evaluate the effect of frequency instability on the probability of distortion of the command, we shall assume as an example that the filter band is equal to

$$\Delta f_F = \Delta f_\tau + \Delta f_{instab}.$$

The first term of the right-hand member, approximately equal to $1/\tau$, is determined by the duration of the elementary signal (the width of its spectrum); the second is determined by the sum total of the frequency instabilities of the transmitter and the heterodyne receiver.

For a binary code

$$\Delta f_{F2} = \frac{1}{\tau_2} + \Delta f_{instab}.$$

For a code with a base m

$$\Delta f_{Fm} = \frac{1}{\tau_m} + \Delta f_{instab}.$$

Transforming the last two expressions:

$$\Delta f_{F_2} = \frac{1}{\tau_2}(1 + \Delta f_{instab}\tau_2) = \frac{1}{\tau_2}(1+b),$$

$$\Delta f_{Fm} = \frac{1}{\tau_2}\left(\frac{1}{\log_2 m} + \Delta f_{instab}\tau_2\right) = \frac{1}{\tau_2}\left(\frac{1}{\log_2 m} + b\right),$$

where

$$b = \Delta f_{instab}\tau_2.$$

The quantity $b = \frac{\Delta f_{instab}}{1/\tau_2}$ shows by what factor the frequency instability exceeds the width of the spectrum which is determined by the duration of the elementary signal.

Then

$$\frac{\Delta f_{F_2}}{\Delta f_{Fm}} = q = \frac{(1+b)\log_2 m}{1 + b\log_2 m}. \quad (20)$$

Substituting the value of q into formula (19) we obtain

$$\frac{P_{Cm}}{P_{C2}} = \frac{m-1}{\log_2 m} \exp\left[\frac{a_2^2(1-\log_2 m)}{4(1+b\log_2 m)}\right]. \quad (21)$$

Using formula (21) one can determine the dependence of the command distortion probability on the frequency stability, characterized by the quantity b. The case where b = 0 corresponds to a high frequency stability ($\Delta f_{instab} = 0$), and the case where $b = \infty$ corresponds to a low frequency stability. Figure 4 shows the results of computations for two values of

m (m = 4, m = 8) and two values of α_2 ($\alpha_2 = 5$, $\alpha_2 = 10$) in curve form. It can be seen from the diagram that the probability of command distortion is significantly dependent on the quantity b; i. e., it depends on the frequency stability.

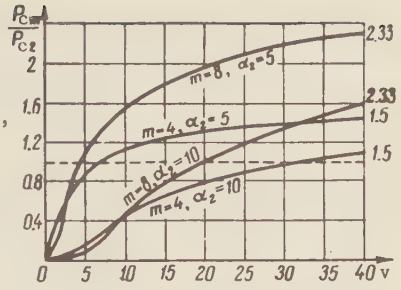

Figure 4

For a very high stability, when b is small, the raising of the code base leads to a decrease in the probability of command distortion in comparison with the probability of distortion for binary double code, while the gain in interference immunity increases with m. With a lowering of the frequency stability, i. e., with an increase in b, the raising of the code base permits one to narrow the transmission band of the filters only at the expense of narrowing the elementary signal spectrum, since that part of it which is determined by the instability remains constant. This unaffected part of the transmission band does not permit one to decrease the level of the interferences to the same degree as for high frequency stability; this in turn leads to an increase in the ratio P_{cm}/P_{c2}. Starting out with certain values of b, corresponding to points of intersection of the curves, the gain in interference immunity will decrease with an increase of the code base. At a certain value of b the ratio P_{cm}/P_{c2} becomes equal to unity and with a further increase in b, approaches $m - 1 / \log_2 m$. The value of b at which the ratio P_{cm}/P_{c2} is equal to unity is a limiting value. At a high value of b, the code with a high base will possess inferior interference immunity characteristics compared to the binary code.

The approximate value of the frequency stability, corresponding to the limiting value of b, can be found in the following manner. Let us assume that m = 4 and $\alpha_2 = 5$. From Figure 4 it can be seen that the limiting value of b corresponding to this case is equal to 7; i. e., $\Delta f_{instab} = 7\frac{1}{\tau_2}$. Let $\tau_2 = 5$ msec; then $\Delta f_{instab} = 7\frac{1}{5 \cdot 10^{-3}} = 1400$ cps. At nonsearch and nontuning communication, the width of the transmission band filter determined by the instability is approximately equal to the quadrupled value of the maximum instability; i. e., $\Delta f_{instab} = 4\Delta f_{max}$. Then $\Delta f_{max} = 350$ cps, and the corresponding relative frequency stability at the maximum frequency of the short-wave communication range $f_{max} = 30 \cdot 10^6$ cps equals

$$\frac{\Delta f_{max}}{f_{max}} = \frac{350}{30 \cdot 10^6} \approx 12 \cdot 10^{-6}.$$

If the frequency stability is higher than $12 \cdot 10^6$ then the transfer to quaternary (i. e., to four-frequency) code leads to a decrease in the command distortion probability. If the frequency stability is below $12 \cdot 10^6$ the transfer to quaternary code leads to an increase in the command distortion probability, and consequently is undesirable. Furthermore, this turns out to be undesirable also at a stability equal to $12 \cdot 10^6$, or somewhat higher, since in that case $P_{c4}/P_{c2} \approx 1$. In actual practice the use of a code with a high base makes sense only in the case when the command distortion probability is decreased by at least a factor of 10; i. e., $P_{c4}/P_{c2} = 0.1$. But then $b \approx 0.8$, $\Delta f_{instab} = 0.8/\tau_2 = 160$ cps, $\Delta f_{max} = 40$ cps, and the relative frequency stability $\frac{\Delta f_{max}}{f_{max}} = 1.33 \cdot 10^{-6}$.

In this manner transfer to quaternary code is justified only for a stability of the order of $1.33 \cdot 10^{-6}$ or higher.

It should be pointed out that the given values of stability are correct for the case where the duration of the elementary signal $\tau_2 = 5$ msec. For other pulse durations, the values of frequency stability at which the raising of the code base becomes justifiable will be different. If, for example, we assume $\tau_2 = 50$ msec it can be easily shown that transfer to quaternary code is reasonable only when the relative stability equals

$$\frac{\Delta f_{max}}{f_{max}} = 1.33 \cdot 10^{-7}.$$

Analogously one can determine the limiting frequency stability for other cases.

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ANALYSIS OF A WIDEBAND LIMITER

M.G. Timishchenko and A.T. Balanov

Investigation is made into a wideband limiter using "ideal" diodes, the internal resistances of which are zero in the forward direction. Relations for computation of currents and voltages in the circuit, as well as for the amplitude and the phase of the first voltage harmonic at the output of the limiter, are derived.

INTRODUCTION

Widespread introduction of the methods of frequency modulation into various areas of radio technology, especially the areas dealing with wideband radio communication, requires that serious attention be paid to the suppression of amplitude modulations in a frequency-modulated wideband signal. Frequency-modulated oscillations, after demodulation, create a signal without distortions only when the signals are free of amplitude changes at the input of the demodulator.

This work deals with one of the problems in limiting a wideband signal — the investigation and development of an engineering procedure for analyzing a wideband limiter.

Physical processes connected with the operations of a limiter over a wide frequency band have been theoretically studied only to a small extent [1, 2], and at the present time no thoroughly developed theory on this matter can be referred to. Similarly there have been no engineering developments for the analysis of wideband limiter circuits. The reason for this apparently lies in the fact that serious difficulties of a mathematical nature appear when a rigorous solution of the problem is attempted. The cause

for this is that limiter circuits contain a combination of reactive elements and elements having strong nonlinear characteristics.

THE LIMITER CIRCUIT

Amplitude limiters used in present day radio transmission apparatus (narrowband), for example, network limiters, possess at frequencies of 70 to 110 Mc (the range of intermediate frequencies of the majority of wideband communication systems) a noticeable inertial character and therefore can not be used for wideband communication apparatus.

Recently, combinations of vacuum tubes, crystal diodes and resonating circuits have been used to limit wideband signals.

The circuit for such a wideband limiter is shown in Figure 1. A single parallel LC_k circuit, serving as a load at the plate of the tube and shunted with a resistance R_{sh}, contains a parallel branch of two crystal diodes, connected in series for DC and in parallel for AC.

A certain bias E_b is applied to the diodes; this controls the magnitude of the voltage in the circuit at which the diodes begin to conduct. If the voltage of oscillation exceeds the bias voltage, then the created diode current will begin to introduce losses into the circuit (shunting the circuit with the diode internal resistance) and thus prevent a further rise in circuit voltage. This phenomenon is the basis of operation of this type of wideband limiter.

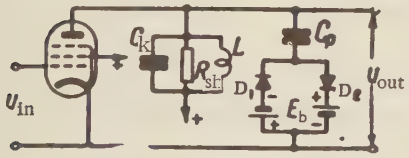

Figure 1

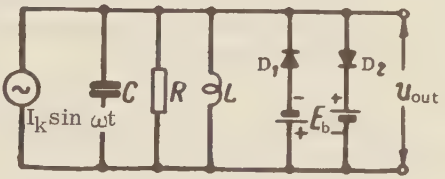

Figure 2

DERIVATION OF EQUATIONS FOR CURRENTS AND VOLTAGES

The equivalent circuit for the limiter is shown in Figure 2. Since the limiter tube (a pentode with a high internal resistance) operates in the linear region, then in the equivalent circuit (Figure 2) it can be represented by a current generator producing $i_k = I_k \sin \omega t$, where $I_k = GmV_{in}$. The graph showing the variation of input (dotted curve) and output (solid curve) voltages with time is shown in Figure 3. (This graph is constructed using the results of the computed example given at the end of this article.) The solution is conducted by an operator method [3,4,5,]. Nonzero initial conditions corresponding to values of voltages equal to $-E_b$ and $+E_b$ on the capacitor C and the current I_{01} and I_{02} in the inductive coil L at the time instants $t_1 = -\psi_1/\omega$ and $t_2 = \psi_2/\omega$ are taken into account. At these time instants, one of the diodes is either nonconducting ($\omega t = \psi_2$) or conducting ($\omega t = -\psi_1$ or $\omega t = \pi - \psi_1$).

If the diodes operated as ideal values then the voltage at the output of such an "ideal" limiter would be at the onset of limitation, constant in its amplitude and equal to the magnitude of the diode bias E_b; i.e., it would have a form very much like a trapezoid. In actuality the diodes are not

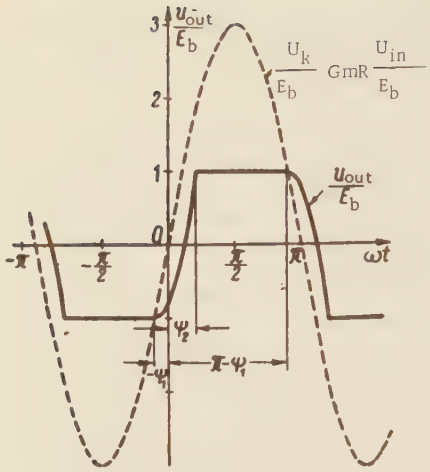

Figure 3

ideal and the voltage has a form which only approximates a trapezoid.

In order to obtain good limitation, as a rule, a high voltage is applied to the diodes and the characteristics of the diode can be represented in the form of a broken straight line.

At the present time our industry produces diodes specially designed for the purposes of limiting; the resistance of these diodes in the forward direction is only a few ohms. Therefore in the analysis of a limiter this small resistance may be neglected, and it may be assumed that the forward resistance of the diodes is equal to zero. No significant errors are introduced into the results of the computations as a result of this assumption, while on the other hand, the procedure for analyzing the limiter is significantly simplified.

Considering the presence of two different regimes in the limiter circuit, we shall look for the solution as follows:

a) for that part of the period when both diodes are nonconducting (first part of the period),

$$-\psi_1 < \omega t < \psi_2;$$

b) for that part of the period when one of the diodes is conducting (the second part of the period),

$$\psi_2 < \omega t < \pi - \psi_1.$$

The initial equations for the first part of the period can be written as follows in operator form:

$$\left.\begin{aligned} U_{c1}(p) &= \frac{I_{c1}(p)}{pC} + \frac{E_3}{p} \\ U_{L1}(p) &= pLI_{L1}(p) - LI_{01} \\ U_{R1}(p) &= RI_{R1}(p), \\ \frac{I_k}{\omega^2 + p^2}(\omega \cos\psi_1 - p \sin\psi_1) &= I_{c1}(p) + I_{L1}(p) + I_{R1}(p) \end{aligned}\right\}, \qquad (1)$$

where $I_{C1}(p)$, $I_{L1}(p)$ and $I_{R1}(p)$ are the transforms for the currents flowing in the branch of capacitor C, inductance coil L, and ohmic resistance R, respectively; I_{01} is the current in the inductance at $\omega t = \pi - \psi_1$.

Solving the system of Equation (1) we find the expression for the current flowing in the branch of resistor R in operator form,

$$I_{R1}(p) = \frac{I_k p(\omega \cos\psi_1 - p \sin\psi_1)}{RC(p^2+\omega^2)\left(p^2 + \frac{p}{RC} + \frac{1}{LC}\right)} + \frac{pE_b}{R\left(p^2 + \frac{p}{RC} + \frac{1}{LC}\right)} - \frac{I_{01}}{RC\left(p^2 + \frac{p}{RC} + \frac{1}{LC}\right)}. \qquad (2)$$

In order to find the original current $i_{R1}(t)$ we will use the inverse Laplace transform method [4]. Considering the retardation theorem and also applying the theory of residues we write

$$I_{R1}(t) = \frac{1}{2\pi i}\int_{C-i\infty}^{C+i\infty} I_{R1}(p)\, e^{p(t+t_1)}\, dp = \sum \text{res}, \tag{3}$$

where Σres is the sum of the residues for the poles of the subintegral function.

Substituting the value of $I_{R1}(p)$, obtained from Equation (2), into Equation (3), we obtain after the proper transformations

$$i_{R1}(t) = \frac{I_k y d}{Z_\omega}\cos(\omega t - \varphi_\omega) - e^{-\tau}\frac{\omega_0}{\omega_p}\left\{I_k d\left[\frac{Z_\psi y}{Z_\omega}\sin(\theta - \varphi_\alpha - \varphi_\psi + \varphi_0) - \right.\right. \tag{4}$$

$$\left.\left. - \sin\theta_1 \sin\psi_1\right] + \frac{E_a}{R}\sin(\theta + \varphi_\alpha)\right\},$$

where

$$\left.\begin{gathered} y = \frac{\omega_0}{\omega};\quad \omega_0 = \frac{1}{\sqrt{LC}};\quad d = \frac{1}{\omega_0 CR} \\ \omega_p = \omega\sqrt{1 - \frac{d^2}{4}};\quad \theta = \frac{\omega_p}{\omega}(\omega t + \psi_1);\quad \tau = \frac{yd}{2}(\omega t + \psi_1) \\ Z_\omega = \sqrt{(y^2-1)^2 + y^2 d^2}; \\ Z_\psi = \sqrt{\cos^2\psi_1 + yd\sin\psi_1\cos\psi_1 + y^2\sin^2\psi_1} \\ \varphi_\omega = \operatorname{arctg}\frac{yd}{y^2-1};\quad \varphi_\alpha = \operatorname{arctg} 2RC\omega_p \\ \varphi_\psi = \operatorname{arctg}\frac{\frac{\omega_p}{\omega}\sin\psi_1}{\cos\psi_1 + \frac{yd}{2}\sin\psi_1};\quad \varphi_0 = \operatorname{arctg}\frac{yd\frac{\omega_p}{\omega}}{1 - y^2 + \frac{y^2 d^2}{2}} \end{gathered}\right\}. \tag{5}$$

Having obtained Equation (4) for the current in the network of the ohmic resistance, the equation for the current in the capacitive branch of the circuit (for the first time period) can easily be found

$$i_{c1}(t) = RC\frac{d}{dt}[i_{R1}(t)] = -\frac{I_k}{Z_\omega}\sin(\omega t - \varphi_\omega) +$$

$$+ e^{-\tau}\frac{\omega_0}{\omega_p}\left[\frac{I_k}{Z_\omega} yZ_\psi \sin(\theta - 2\varphi_\alpha - \varphi_\psi + \varphi_0) - \right.$$

$$\left. - I_k \sin(\theta - \varphi_\alpha)\sin\psi_1 + \frac{E_b}{Rd}\sin\theta\right]. \tag{6}$$

In order to determine the current in the inductive branch we shall use Kirchoff's first law

$$i_{L1}(t) = I_k \sin\omega t - i_{R1}(t) - i_{c1}(t) = \frac{I_k}{Z_\omega} y^2 \sin(\omega t - \varphi_\omega) +$$

$$+ e^{-\tau}\frac{\omega_0}{\omega_p}\left[\frac{I_k}{Z_\omega} yZ_\psi \sin(\theta - \varphi_\psi + \varphi_0) - I_k \sin\psi_1 \sin(\theta + \varphi_\alpha) + \right.$$

$$\left. + \frac{E_b}{Rd}\sin(\theta + 2\varphi_\alpha)\right]. \tag{7}$$

Taking into account Equation (4) the expression for the voltage at the output of the limiter (in the circuit) may be easily obtained

$$U_{R1}(t) = \frac{I_k}{\omega C Z_\omega}\cos(\omega t - \varphi_\omega) -$$

$$- e^{-\tau}\frac{\omega_0}{\omega_p}\left\{\frac{I_k}{\omega_0 C}\left[y\frac{Z_\psi}{Z_\omega}\sin(\theta-\varphi_a-\varphi_\psi+\varphi_0)-\sin\theta\sin\psi_1\right]+\right.$$
$$\left.+E_b\sin(\theta+\varphi_a)\right\}. \tag{8}$$

Dividing the left- and right-hand members of Equation (8) by E_b and grouping the terms containing cos θ and sin θ we finally obtain

$$\frac{U_{P1}(t)}{E_b}=M\cos(\omega t-\varphi_\omega)-N\,e^{-\tau}\sin(\theta+\varphi_N), \tag{9}$$

where

$$M=\frac{yd}{Z_\omega}x, \qquad x=\frac{I_k R}{E_b}=SR\,\frac{V_{in}}{E_b}, \tag{10}$$

$$N=\frac{1+M\frac{\omega_0}{\omega_p}Z_\psi\sin(\varphi_0-\varphi_a-\varphi_\psi)}{\sin\varphi_N}, \tag{11}$$

$$\varphi_N=\operatorname{arctg}\frac{1+M\frac{\omega_0}{\omega_p}Z_\psi\sin(\varphi_0-\varphi_a-\varphi_\psi)}{\frac{\omega_0}{\omega_p}\left[MZ_\psi\cos(\varphi_0-\varphi_a-\varphi_\psi)-xd\sin\psi_1+\frac{d}{2}\right]}. \tag{12}$$

The resistance R is equivalent to the four resistances R_{sh}, R_i, R_e and R_c connected in parallel and is equal to

$$R=\frac{1}{\frac{1}{R_{sh}}+\frac{1}{R_i}+\frac{1}{R_e}+\frac{1}{R_c}}; \tag{13}$$

here $R_e = L/Cr$ is the equivalent resistance of the unshunted circuit at resonance; R_i is the internal resistance of the limiter tube; R_c is the input resistance of the next stage; r is the resistance of the losses of the unshunted circuit.

For the second part of the period ($\psi_2 < \omega t < \pi - \psi_1$) the voltage across the circuit $u_{R2}(t)$ is equal to the voltage of the bias E_b (since it was assumed that the internal resistance of the diodes in the forward direction is zero); i. e.,

$$\frac{U_{F2}(t)}{E_b}=1. \tag{14}$$

The current in the inductive branch $i_{L2}(t)$ is found from the condition

$$E_b=L\frac{d}{dt}[i_{L2}(t)],$$

from where, using the Riemann-Mellin theorem and the theorem of residues, we finally obtain

$$i_{L2}(t)=I_k\sin\psi_1-\frac{E_b}{R}+\frac{E_b y}{Rd}(\omega t-\pi+\psi_1). \tag{15}$$

Since the circuit voltage during the second part of the period is constant, the current in the capacitive branch will be equal to zero.

It can be seen from Equations (4), (5), (6), (7), (8) and (15) that the values of currents and voltages are expressed through the R, L and C parameters of the circuit, the magnitude of the relative detuning y, and the cutoff angles ψ_1 and ψ_2. On the other hand, the magnitude of the cutoff angles ψ_1 and ψ_2 are determined by the parameters and the operating regime

of the entire limiter circuit in general. Therefore, in order to analyze the limiter it becomes necessary to find the relation between the unknown angles ψ_1 and ψ_2 and the parameters of the circuit and its operating regime.

RELATIONS FOR COMPUTING CUTOFF ANGLES ψ_1 AND ψ_2

Since there are two unknown angles (ψ_1 and ψ_2), we must develop a system of two equations in order to solve for these quantities. For this purpose we will utilize the following limiting conditions:

a) the condition of continuity of current in the inductance at the junction of the two segments of the period, corresponding to $\omega t = \psi_2$

$$[i_{L1}(\omega t)]_{\omega t=\psi_2} = [i_{L2}(\omega t)]_{\omega t=\psi_2}; \tag{16}$$

b) the condition of continuity of voltage across the capacitor

$$[u_{c1}(\omega t)]_{\omega t=\psi_2} = E_b. \tag{17}$$

Substituting $\omega t = \psi_2$ into Equations (7) and (15) and taking condition (16) into account, we obtain the first equation for the determination of angles ψ_1 and ψ_2

$$x(W \sin \psi_1 - B \cos \psi_1) = A, \tag{18}$$

where

$$\left.\begin{aligned} W &= 1 + \frac{y^2}{Z_\omega} \cos(\beta - \varphi_\omega) + \\ &\quad + e^{-\eta} \frac{\omega_0}{\omega_\rho} \left[\sin(k + \varphi_a) - \frac{y^2}{Z_\omega} \sin(k + \varphi_0 - \varphi_a) \right] \\ B &= \frac{y}{Z_\omega} \left[y \sin(\beta - \varphi_\omega) + e^{-\eta} \frac{\omega_0}{\omega_\rho} \sin(k + \varphi_0) \right. \\ A &= 1 + \frac{y}{d}(\pi - \beta) + e^{-\eta} \frac{\omega_0}{\omega_\rho} \frac{1}{d} \sin(k + 2\varphi_a) \end{aligned}\right\}, \tag{19}$$

$$k = \frac{\omega_\rho}{\omega} \beta, \qquad \eta = \frac{yd}{2} \beta, \quad \beta = \psi_1 + \psi_2. \tag{20}$$

Using condition (17) and Equation (8) we obtain the second equation required for determination of angles ψ_1 and ψ_2

$$x(F \sin \psi_1 + T \cos \psi_1) = H, \tag{21}$$

where

$$\left.\begin{aligned} F &= \frac{d}{y} B \\ T &= \frac{yd}{Z_\omega} \left[\cos(\beta - \varphi_\omega) - e^{-\eta} \frac{\omega_0}{\omega_\rho} \sin(k + \varphi_0 - \varphi_a) \right] \\ H &= 1 + \frac{\omega_0}{\omega_\rho} e^{-\eta} \sin(k + \varphi_a) \end{aligned}\right\}. \tag{22}$$

Solving Equations (18) and (21) simultaneously we obtain the ratio for the determination of the cutoff angles ψ_1 and ψ_2 and the parameter x:

$$\operatorname{tg} \psi_1 = -\frac{BH + AT}{WH - AF}, \tag{23}$$

$$\psi_2 = \beta - \psi_1, \tag{24}$$

$$x = \frac{A}{W \sin \psi_1 - B \cos \psi_1}. \qquad (25)$$

It can be seen from (19) and (22) that it is necessary to know the attenuation of the circuit d, the magnitude of the relative detuning of the circuit y, and the sum of the cutoff angles $\psi_1 + \psi_2 = \beta$ in order to be able to compute the coefficients W, B, A, H, T and F.

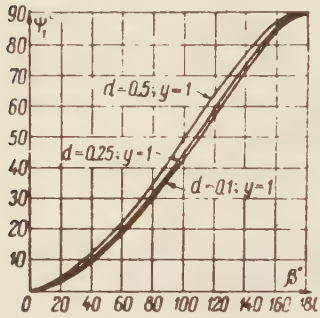

Figure 4

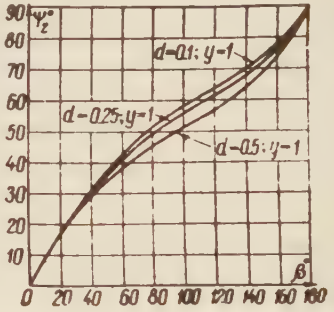

Figure 5

The attenuation of the circuit d, determined by the required wide bandwidth of the limiter, and the magnitude of the relative detuning of the circuit y are given during the computations. The unknown is the summed cutoff angle β. However, angle β may vary only within the limits of 0-180°. Therefore we will use the following procedure for analyzing the limiter. Assuming the values of β we will find the magnitudes of the coefficients W, B, A, H, T and F for constant values of d and y. Then, using Equations

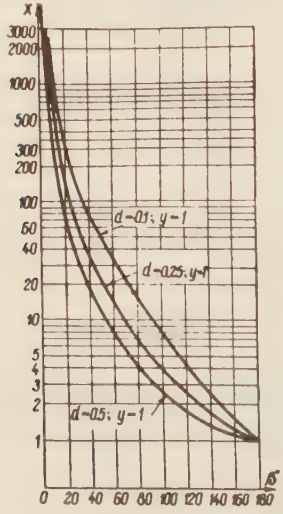

Figure 6

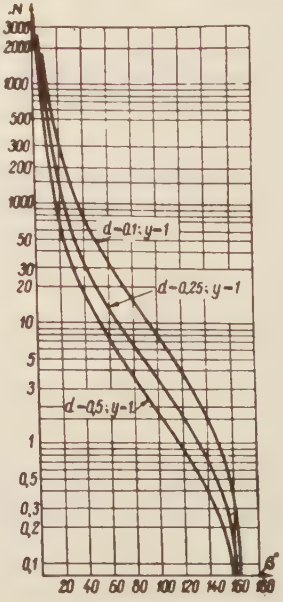

Figure 7

(23) and (24) we find the relation between ψ_1 and $\psi_2 = \varphi(\beta)$, which is shown for some values d and y in Figures 4 and 5. In this manner to each value of the angles ψ_1 and ψ_2 there corresponds a completely determined value of the coefficients W, B, A, H, T and F.

Turning our attention to Equation (25) we find the relation $x = \varphi(\psi_1, \psi_2)$ or $x = \varphi(\beta)$. For several values of d and y this relation is shown in Figure 6.

In order to compute the relation $u_{R_1} = \varphi(U_{in})$ it is convenient to make use of Equation (9) where, for computational simplicity, it is useful to first compute the dependences $N = \varphi(\beta)$ and $\varphi_N = \varphi(\beta)$. These dependencies, for the same values of d and y as shown in Figures 5 and 6, are given in Figures 7 and 8.

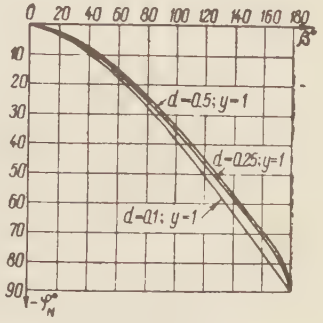

Figure 8

Thus, it can be seen from the presented limiter analysis that if the sum of the angles $\psi_1 + \psi_2 = \beta$ are given as functions of the required degree of limiting, and the parameters d and y are given as functions of the bandwidth, the unknowns x, ψ_1 and ψ_2 can be determined in a direct manner. At the end of this article a concrete numerical example is presented to illustrate the peculiarities of wideband limiter analysis.

RELATIONS FOR COMPUTING AMPLITUDE AND PHASE OF THE LIMITER FIRST HARMONIC OUTPUT VOLTAGE

Since the voltage from the output of the limiter is fed to the grid of the tube in the next stage, the plate load of which is the circuit, the higher harmonics, except for the first, can be neglected since they will be either completely filtered out, or decreased to such a degree that their effect can be disregarded. Therefore it makes sense to find the equation for the amplitude and phase of only the first harmonic of the voltage at the output of the limiter.

The expression for the relative voltage at the output of the limiter, in the general form, can be written as follows:

$$\frac{u_{out}(t)}{E_b} = \frac{V_1}{E_b} \sin(\omega t + \varphi), \tag{26}$$

where

$$\frac{V_1}{E_b} = \frac{(b_1)}{E_b} \frac{1}{\cos\varphi} \tag{27}$$

is the relative value of the amplitude and

$$\varphi = \operatorname{arctg}\left(\frac{b_1}{E_b}\right) \Big/ \left(\frac{a_1}{E_b}\right) \tag{28}$$

is the value of the phase angle of the first harmonic of the output voltage

The relations for computing the coefficients a_1/E_b and b_1/E_b can be determined by expressing Equations (9) and (14) as a Fourier series which describes the law of changes of the voltage in the circuit of a limiter:

$$\frac{a_1}{E_b} = \frac{2}{\pi}\left[\int_{-\psi_1}^{\psi_2} \frac{u_{\Sigma 1}(\omega t)}{E_b}\cos\omega t\, d\,\omega t + \int_{\psi_2}^{\pi-\psi_1} \frac{u_{\Sigma 2}(\omega t)}{E_b}\cos\omega t\, d\,\omega t\right] =$$

$$= \frac{1}{\pi}\left\{M[\beta\cos\varphi_\omega + \sin\beta\cos(\psi_2 - \varphi_1 + \varphi_\omega)] + 2(\sin\psi_1 - \sin\psi_2) - \right.$$

$$- N\left[\frac{1}{Z_1}\sin(\psi_1 + \varphi_N + \varphi_1) - \frac{1}{Z_2}\sin(\psi_1 - \varphi_N - \varphi_2)\right] +$$

$$\left. + Ne^{-\gamma}\left[\frac{1}{Z_1}\sin(k - \psi_2 + \varphi_N + \varphi_1) + \frac{1}{Z_2}\sin(k + \psi_2 + \varphi_N + \varphi_2)\right]\right\}; \qquad (29)$$

$$\frac{b_1}{E_b} = \frac{2}{\pi}\left[\int_{-\psi_1}^{\psi_2} \frac{u_{\Sigma 1}(\omega t)}{E_b}\sin\omega t\, d\,\omega t + \int_{\psi_2}^{\pi-\psi_1} \frac{u_{\Sigma 2}(\omega t)}{E_b}\sin\omega t\, d\,\omega t\right] =$$

$$= \frac{1}{\pi}\left\{M[\beta\sin\varphi_\omega + \sin\beta\sin(\psi_2 - \psi_1 - \varphi_\omega)] + 2(\cos\psi_1 + \cos\psi_2) - \right.$$

$$- N\left[\frac{1}{Z_1}\cos(\psi_1 + \varphi_N + \varphi_1) - \frac{1}{Z_2}\cos(\varphi_1 - \varphi_N - \varphi_2)\right] +$$

$$\left. + Ne^{-\gamma}\left[\frac{1}{Z_1}\cos(k - \psi_2 + \varphi_N + \varphi_1) - \frac{1}{Z_2}\cos(k + \psi_2 + \varphi_N + \varphi_2)\right]\right\}. \qquad (30)$$

Here

$$\left.\begin{aligned} Z_1 &= \sqrt{\left(\frac{yd}{2}\right)^2 + (y_p - 1)^2}\ ;\ \varphi_1 = \operatorname{arctg}\frac{y_p - 1}{\frac{yd}{2}}; \\ Z_2 &= \sqrt{\left(\frac{yd}{2}\right)^2 + (y_p + 1)^2}\ ;\ \varphi_2 = \operatorname{arctg}\frac{y_p + 1}{\frac{yd}{2}};\ y_p = \frac{\omega_p}{\omega} \end{aligned}\right\}. \qquad (31)$$

As can be seen from Equations (29), (30) and (31), the coefficients of resolution a_1/E_b and b_1/E_b can be directly expressed by means of the known (given for analysis purposes) parameters of the limiter x, ψ_1, ψ_2, d and y.

In this manner, having calculated coefficients a_1/E_b and b_1/E_b and by using Equations (27) and (28), we can determine the amplitude and phase characteristic of the limiter.

EXAMPLE

It is desired to determine the basic elements of the circuit and also to find the voltage form at the output of the limiter, if a sinusoidal voltage $u_{in} = V_{in}\sin\omega_0 t$ with an amplitude $V_{in} = 1.5$ volts, and a frequency of $f_0 = 70$ Mc is applied to the input. The transmission bandwidth of the limiter must be $2\Delta f = 40$ Mc and the output voltage $V_{out} = 3$ volts.

Solution.

In order to determine the operating regime of the limiter it becomes necessary to compute parameter x, using Equation (10). Since, for the given amplitude of input voltage V_{in}, for a constant magnitude of the limiter output voltage ($V_{out} = E_b$), it is desired to achieve the best possible limiting

of the wideband signal, i.e., to obtain the highest possible value of the parameter x, it is expedient to use a tube with a high transconductance and the least possible input capacitance in the limiting stage.

A suitable tube [7] is a high frequency pentode of the bantam series of the type 6Zh11P, which has a transconductance Gm = 28 ma/v, an output capacitance C_{out} = 3.5 $\mu\mu f$, and an internal resistance of R_i = 34 kilohms.

Starting out with the given value of the transmission band we determine the attenuation of the circuit [6]

$$d = \frac{2\Delta\omega}{\omega} \cdot \frac{1+0.5\,\frac{\Delta\omega}{\omega}}{1+\frac{\Delta\omega}{\omega}} = 0.5.$$

Selecing the sum-total capacitance of the circuit (this capacitance is equal to the sum of the output capacitance of the tube of the limiter, the input capacitance of the next stage, the input capacitance of the diodes, and the capacitance of the circuit chassis), equal to 8 $\mu\mu f$, we go on to find the equivalent resistance which must be used for shunting the circuit to attain the given (40 Mc) transmission band,

$$R = \frac{1}{\omega_0 C d} = 570 \text{ ohms}.$$

The inductance of the circuit equals

$$L = \frac{1}{\omega_0^2 C} = 0.646\,\mu H.$$

In this manner using the known data we obtain

$$x = SR\,\frac{V_{in}}{E_b} = 7.98.$$

Having determined the parameter x, from the graph in Figure 6 we find the value of the summed cutoff angle β to be 60°. From graphs in Figures 4 and 5 we find $\psi_1 = 21°30'$, $\psi_2 = 38°30'$. And finally, from graphs in Figures 7 and 8 we find the value N = 7.6 and $\varphi_N = -14°40'$.

Turning now to Equation (9) and substituting into it the known quantities, we obtain

$$\frac{u_{p1}(\omega t)}{E_b} = 7.98 \sin \omega t - 7.6\, e^{-0.25\left(\omega t + \frac{\pi}{180°}\,21°30'\right)} \sin(0.968\omega t + 6°08').$$

On the basis of the results of the computations, a graph showing the dependence of output voltage on time has been constructed in Figure 3. Taking various values of input voltage it becomes easy to compute the amplitude characteristic of the limiter.

CONCLUSIONS

The obtained Equations (4), (6), (7), (8) and (14) permit one to calculate the voltages and currents in the wideband limiter circuit with a consideration for transient phenomena brought about by the presence of the diodes, which conduct only at the definite value of circuit voltages equal to the magnitude of bias voltage of the diodes E_b.

Conditions (21) and (22) created the possibility of obtaining a system of two transcendental Equations (23) and (25) which permit the obtaining of relations for the computation of cutoff angles ψ_1 (27) and ψ_2 (28).

On the basis of Equation (25), considering (10), it is possible, at different operating regimes of the limiter, to determine the limiting permissible tube current at a required wide bandwidth, i.e., at a given d. In this manner the value of parameter x permits the selection of the optimum tube type as a function of the definite requirements for the limiter. Knowing the parameters of the tube one can determine the greatest depth of limiting (β) at a given diode bias voltage, E_b, determining the level of the signal at the output of the limiter.

Results computed from the derived formulas were compared with experimental values. A good agreement of experimental and computed values leads to a hope that the presented procedure will find application in engineering practice.

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AC TRANSISTOR AMPLIFIER USING DIRECT COUPLING

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The static stability of an amplifier, taking into account the effects of changes in I_{c0}, α and U_{be}, is investigated. The amplifier circuit is shown and the procedure for analysis is presented.

Transistorized AC amplifiers with direct coupling are used in automatic control systems as preamplifiers, and as variable voltage amplifiers in operational amplifiers with automatic zero correction. The absence of transformers and capacitors in the coupling networks makes this type of amplifier especially suitable for microminiaturization applications. However, widespread application of amplifiers using direct coupling is hampered by the relative complexity of selection of a static regime for the amplifier, which is sufficiently stable during changes in ambient temperature and during changes in transistor parameters.

Material presented in the literature dealing with the design of direct-coupled amplifiers is basically of a qualitative nature. Works [1, 2] which present a procedure for analysis do not encompass all circuit variations, and during analysis they consider only the changes in the zero collector current I_{c0}. As will be shown below this is not adequate, especially in the case of silicon transistors.

The analysis of static stability of the direct-coupled amplifier which takes into account changes in the transistor parameters — zero collector current I_{c0}, the current coefficient of amplification α, and the base-emitter voltage U_{be} — are discussed. The procedure for designing an amplifier with DC feedback is also given.

STATIC STABILITY

The parameters of the transistors change with the temperature of the surrounding medium. This leads to a change of the static regime of the amplifier and may impair its operating ability. The basic effect on the changes of the static regime of the amplifier is brought about by the changes in the zero collector current I_{c0}, the current coefficient of amplification α, and the base-emitter voltage U_{be}. Thus in the temperature range from +20° to 60°C the factor β-α/1-α increases by 25 to 30% for germanium and silicon transistors P13-P15 and P101-P103; in the temperature range of +20°C to-60°C, β decreases by 30-50% for transistors P13-P15, and by 50% or more for transistors P101-P103. With an increase in temperature,

I_{c0} increases by approximately a factor of two for each 10°C in the case of transistors P13-P15; for transistors P101-P103 at a temperature of +60°C it does not exceed 1 μa. In the temperature interval of +20°C to 60°C U_{be} decreases by 0.03 to 0.1 volts and at negative temperatures it increases by 0.15 to 0.2 volts in the case of transistors P101-P103.

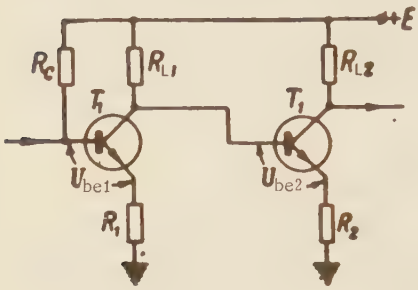

Figure 1

In view of space limitations, we will evaluate the effect of changes in the indicated parameters on the static regime of the amplifier and also the order of derivation of the expressions determining the stability of the static regime, using as an example a two-stage amplifier, the circuit of which is shown in Figure 1.

In order to obtain the expression for the collector current of the second stage of the amplifier let us, in accordance with the circuit, write the following equation:

$$E - (I_{c1} + I_{b2}) R_{L1} - U_{be2} - R_2 (I_{c2} + I_{b2}) = 0, \tag{1}$$

where I_{c1} is the current of the collector of the first transistor; I_{c2} and I_{b2} are the collector and base currents of the second transistor respectively.

Substituting into expression (1) the value

$$I_{b2} = \frac{I_{c2}(1 - \alpha_2) - I_{c02}}{\alpha_2}$$

and making simple transformations we obtain

$$I_{c2} = \frac{E\alpha_2}{R_2 + R_{L1}(1 - \alpha_2)} - \frac{U_{be2}\alpha_2}{R_2 + R_{L1}(1 - \alpha_2)} - \frac{I_{c1}R_{L1}\alpha_2}{R_2 + R_{L1}(1 - \alpha_2)} + \frac{I_{\kappa 02}(R_{L1} + R_2)}{R_2 + R_{L1}(1 - \alpha_2)},$$

or

$$I_{c2} = \frac{E(S_2 - 1)}{R_{L1}} - \frac{U_{be2}(S_2 - 1)}{R_{L1}} - I_{c1}(S_2 - 1) + I_{c02}S_2, \tag{2}$$

where $S_2 = \frac{R_{L1} + R_2}{R_2 + R_{L1}(1 - \alpha_2)} = \frac{\partial I_{c2}}{\partial I_{c02}}$ is the coefficient of stability, determining the degree of the effect of I_{c0} on the collector current. To determine the collector current of the first stage we set up the equation

$$E - I_{b1}R_c - U_{be1} - R_1(I_{b1} + I_{c1}) = 0. \tag{3}$$

Substituting into (3) the value of the base current, expressed in terms of the collector current, similarly to the last case, and making analogous transformations, we obtain

$$I_{c1} = \frac{E}{R_c}(S_1 - 1) - \frac{U_{be1}}{R_c}(S_1 - 1) + I_{c01}S_1, \tag{4}$$

where

$$S_1 = (R_c + R_1)[R_1 + R_c(1 - \alpha_1)].$$

After substituting expression (4) into (2) the collector current of the second stage will be written as

$$I_{c2} = \frac{E}{R_{L1}}(S_2 - 1) - \frac{U_{be2}}{R_{L1}}(S_2 - 1) - \frac{E}{R_c}(S_1 - 1)(S_2 - 1) + \\ + \frac{U_{be1}}{R_c}(S_1 - 1)(S_2 - 1) - I_{c01}(S_2 - 1)S_1 + I_{c02}S_2. \quad (5)$$

Assuming that the supply voltage E and the resistance of the bias networks are constant, the complete differential of the collector current of the second stage, in general form, can be written as

$$dI_{c2} = \frac{\partial I_{c2}}{\partial I_{c01}} dI_{c01} + \frac{\partial I_{c2}}{\partial I_{c02}} dI_{c02} + \frac{\partial I_{c2}}{\partial \alpha_1} d\alpha_1 + \frac{\partial I_{c2}}{\partial \alpha_2} d\alpha_2 + \\ + \frac{\partial I_{c2}}{\partial U_{be1}} dU_{be1} + \frac{\partial I_{c2}}{\partial U_{be2}} dU_{be2}.$$

Using finite increments

$$\Delta I_{c2} \approx C_1 \Delta I_{c01} + C_2 \Delta I_{c02} + A_1 \Delta \alpha_1 + A_2 \Delta \alpha_2 + H_1 \Delta U_{be1} + H_2 \Delta U_{be2}$$

For expression (5)

$$C_1 \Delta I_{c01} = -S_1(S_2 - 1)\Delta I_{c01} \quad (5'); \quad C_2 \Delta I_{c02} = S_2 \Delta I_{c02} \quad (5'');$$

$$A_1 \Delta \alpha_1 = C_1 \frac{I_{c1} \Delta \alpha_1}{\alpha_1 + \Delta \alpha_1} \quad (5'''); \qquad A_2 \Delta \alpha_2 = C_2 \frac{I_{c2} \Delta \alpha_2}{\alpha_2 + \Delta \alpha_2} \quad (5''''');$$

$$H_1 \Delta U_{be1} = -\frac{(S_1 - 1)(S_2 - 1)}{R_c} \Delta U_{be1} \quad (5'''''); \quad H_2 \Delta_{be2} = \frac{(S_2 - 1)\Delta U_{be2}}{R_{L1}} \quad (5''''''').$$

In this manner we may write

$$\Delta I_{c2} \approx \frac{I_{c2} \Delta \alpha_2}{\alpha_2 + \Delta \alpha_2} S_2 - S_1(S_2 - 1)\frac{I_{c1} \Delta \alpha_1}{\alpha_1 + \Delta \alpha_1} + S_2 \Delta I_{c02} - S_1(S_2 - 1)\Delta I_{c01} + \\ + \frac{S_2 - 1}{R_{L1}} \Delta U_{be2} - \frac{(S_1 - 1)(S_2 - 1)}{R_c} \Delta U_{be1} \quad (6)$$

where $\Delta\alpha$, ΔI_{c0}, ΔU_{be} are the magnitudes of change of the current coefficient of amplification, zero collector current and the voltage of the base-emitter respectively, due to a change in temperature.

The relative change of the second-stage collector current is determined in the following manner:

$$\delta I_{c2} \approx \Delta I_{c2} / I_{c2}. \quad (7)$$

For a decrease in temperature α and I_{c0} decrease and U_{be} increases, i.e., the direction of the variations changes, and this must be considered during the computation of collector current variations.

Analysis of expressions (6) and (7) permits us to draw the following conclusions:

1. Expressions (6) and (7) allow determination of the absolute and relative change of the collector current of the second stage when the changes in α, I_{c0} and U_{be} are known, i.e., permit the evaluation of static stability.
2. For a definite relation between the currents, coefficients of stability and the magnitudes of changes in α, I_{c0} and U_{be} the change in collector current of the second stage ΔI_{c2} may be equal to zero.
3. The effect of changes in α, I_{c0} and U_{be} of the preceding stage will increase by a factor of $(S_1 - 1)$.

In order to check the correctness of the obtained expressions (6) and (7) and the necessity of considering the changes in α and U_{be} we present the

calculated results and the experimental check for a two-stage amplifier assembled with transistors P102 with the following parameters: E = 4 v, R_0 = 150 kilohms, $R_1 = R_2$ = 500 ohms, R_{L1} = 6.2 kilohms, R_{L2} = 3 kilohms, I_{c1} = 0.42 ma, I_{c2} = 0.86 ma, α_1 = 0.953, α_2 = 0.857.

At a temperature of +60°C, $\Delta\alpha_1$ = 0.01; $\Delta\alpha_2$ = 0.033, ΔU_{be1} = 0.05 v, ΔU_{be2} = 0.05 v, ΔI_{c2} (computed) = 254 μa, ΔI_{c2} (experimental) = 230 μa.

With this

$$C_1 \Delta I_{c01} + C_2 \Delta I_{c02} = -15 \quad \mu a;$$

$$H_1 \Delta U_{be1} + H_2 \Delta K_{be2} = 38 \quad \mu a; \quad A_2 \Delta \alpha_1 + A_2 \Delta \alpha_2 = -277 \quad \mu a.$$

As can be seen from the presented data the calculated and experimental results agree sufficiently well and the incremental growth in collector current basically is determined by the components which depend upon changes in α. It follows from here that in computing the static stability of the amplifier the change in the current coefficient of amplification α must be considered, especially for silicon-type transistors.

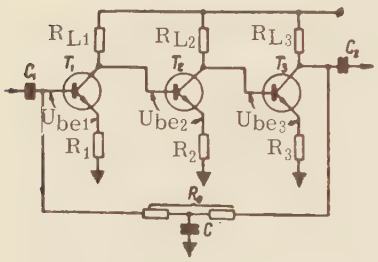

Figure 2

In multistage amplifiers, in order to attain a definite stability of the static regime, aside from local current and voltage feedback, a negative voltage feedback for the DC component is used, encompassing three stages, as shown in Figure 2. There is no connection for AC due to the shunting effect of capacitance C.

In order to determine the stability of the static operating point of the amplifier, the circuit of which is shown in Figure 2, we make use of Equation (8)

$$E - (I_{c3} + I_{b1}) R_{L3} - I_{b1} R_0 - U_{be1} - R_1 (I_{c1} + I_{b1}) = 0, \tag{8}$$

where I_{c3} is the collector current of the third transistor; I_{b1} and I_{c1} the currents of the base and the collector respectively of the first transistor.

Using the procedure described above we find the expressions for collector currents of the amplifier stages:

$$I_{c1} = I_1 (S_1 - 1) - m I_{c3} (S_1 - 1) + I_{c01} S_1, \tag{9}$$

$$I_{c2} = I_2 (S_2 - 1) - (S_2 - 1)[I_1 (S_1 - 1) - m I_{c3} (S_1 - 1) + I_{c01} S_1] + I_{c02} S_2, \tag{10}$$

$$I_{c3} = \frac{1}{Q} [I_3 (S_3 - 1) - I_2 (S_3 - 1)(S_2 - 1) + I_1 (S_1 - 1)(S_2 - 1)(S_3 - 1) + I_{c01} (S_3 - 1)(S_2 - 1) S_1 - I_{c02} (S_3 - 1) S_2 + I_{c03} S_3], \tag{11}$$

where

$$I_1 = \frac{E - U_{be1}}{R_0 + R_{L2}}; \quad I_2 = \frac{E - U_{be2}}{R_{L1}}; \quad I_3 = \frac{E - U_{be3}}{R_{L2}};$$

$$S_1 = \frac{R_{L3} + R_0 + R_1}{R_1 + (R_{L3} + R_0)(1 - \alpha_1)}; \tag{12}$$

$$S_2 = \frac{R_{L1} + R_2}{R_2 + R_{L1}(1 - \alpha_2)}; \tag{13}$$

$$S_3 = \frac{R_{L2} + R_3}{R_3 + R_{L2}(1 - \alpha_3)}; \tag{14}$$

$$m = \frac{R_{L3}}{R_{L3} + R_0}; \tag{15}$$

$$Q = 1 + m(S_1 - 1)(S_2 - 1)(S_3 - 1).$$

The value of change in collector current of the third stage due to the changing of α_1, α_2, α_3, I_{co1}, I_{co2}, I_{co3}, U_{be1}, U_{be2}, and U_{be3} will be expressed in the following manner:

$$\Delta I_{c3} \approx C_1 \Delta I_{co1} + C_2 \Delta I_{co2} + C_3 \Delta I_{co3} + A_1 \Delta\alpha_1 + A_2 \Delta\alpha_2 + A_3 \Delta\alpha_3 + \\ + H_1 \Delta U_{be1} + H_2 \Delta U_{be2} + H_3 \Delta U_{be3}. \tag{16}$$

Here

$$C \approx \frac{S_1}{m(S_1 - 1)}; \tag{16'}$$

$$C_2 \approx -\frac{S_2}{m(S_1 - 1)(S_2 - 1)}; \tag{16''}$$

$$C_3 \approx \frac{S_3}{m(S_1 - 1)(S_2 - 1)(S_3 - 1)}; \tag{16'''}$$

$$A_1 = \frac{C_1 I_{c1}}{\alpha_1 + \Delta\alpha_1}; \tag{17'}$$

$$A_2 = \frac{C_2 I_{c2}}{\alpha_2 + \Delta\alpha_2}; \tag{17''}$$

$$A_3 = \frac{C_3 I_{c3}}{\alpha_3 + \Delta\alpha_3} \approx 0; \tag{17'''}$$

$$H_1 = 1/R_{L3}; \tag{18'}$$

$$H_2 = 1/m(S - 1)R_{L1}; \tag{18''}$$

$$H_3 = 1/m(S_1 - 1)(S_2 - 1)R_{L2}. \tag{18'''}$$

If silicon transistors are used in the amplifier, the magnitude of the change of the collector current can be determined by the following approximate expressions:

$$\Delta I_{c3} \approx \frac{I_{c1} S_1 \Delta\alpha_1}{m(S_1 - 1)(\alpha_1 + \Delta\alpha_1)} - \frac{I_{c2} S_2 \Delta\alpha_2}{m(S_1 - 1)(S_2 - 1)(\alpha_2 + \Delta\alpha_2)}. \tag{19}$$

The relative change of the collector current of the third stage

$$\delta I_{c3} \approx \Delta I_{c3} / I_{c3}. \tag{20}$$

Expressions (19) and (20) permit the determination of the change in collector current of the third stage and, in that manner, the determination of the stability of the static regime of the three-stage amplifier with DC feedback. For a more accurate determination of the stability of the static regime of the amplifier it becomes necessary to use expression (16).

AMPLIFIER DESIGN PROCEDURE

During the design of an amplifier the given technical requirements usually are the supply voltage E, maximum amplitude of the output voltage U_{out} and the voltage coefficient of amplification K_u. Computations must begin with the last stage. In the amplifier with negative feedback, the circuit of which is shown in Figure 2, the feedback resistance shunts load R_{L3}; however, since R_0 is usually much larger than R_{L3} the former may be

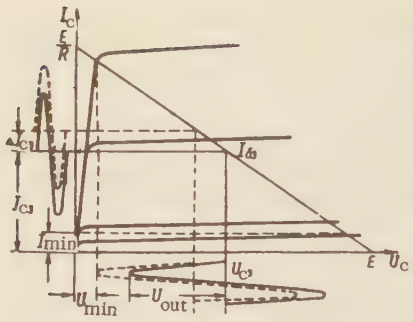

Figure 3

neglected in the computation of the last stage.

The DC load line is plotted on a family of collector characteristics (Figure 3), with respect to constant current [4]. In our case R_{L3} is the AC load.

If a feedback resistance R_3 is present in the emitter circuit, the load line is plotted with a slope whose magnitude is determined by the resistance $R=R_{L3}+R_3$ [3].

The operating point of the stage is selected in such a manner that even when it is displaced by changes in temperature (the DC component of the collector current I_{C3} is changed by the value ΔI_{C3}), the instantaneous value of the collector current will be $\geqslant I_{min}$, and the instantaneous value of the output voltage will be $\geqslant U_{min}$, since in the region of small currents I_{min} the transfer characteristic becomes nonlinear and U_{min} — the initial nonlinear sector of collector characteristics.

Taking the indicated limitations into account we determine the collector current I_{C3}, the base current I_{b3}, the current amplification coefficient $\beta_3(\alpha_3)$ and also the permissible value of the change in collector current ΔI_{C3} during changes in temperature.

The voltage drop across the feedback resistance is determined by the expression

$$U_{c3} = R_3 (I_{b3} + I_{c3}). \tag{21}$$

Using the output characteristics, for given values of collector voltage U_{C3} and base current I_{b3} we find the base-emitter voltage U_{be3}.

Let us determine the static operating collector voltage of the preceding stage

$$U_{c2} = U_3 + U_{be3}. \tag{22}$$

The input resistance of the third stage equals

$$R_{in3} \approx R_3/(1-\alpha_3). \tag{23}$$

We assume the voltage drop for the feedback resistance R_2 in the emitter circuit of the second transistor to be U_2.

Let us determine the resistance R_{L2} using formula [4]

$$R_{L2} = R_{in3} \frac{E - 2U_{c2} + U_{cmin2}}{U_{c2} - U_{cmin2} + R_{in3} I_{cmin2}}, \tag{24}$$

where

$$U_{cmin2} = U_2 + U_{min2}. \tag{25}$$

We determine the second transistor collector current

$$I_{c2} = \frac{E - U_{c2}}{R_{L2}} - I_{b3}. \tag{26}$$

We determine the base current I_{b2} using the collector characteristics for given values of U_{C2} and I_{C2}.

The feedback resistance of the second stage is found from the relation

$$R_2 = U_2/(I_{c2} + I_{b2}). \quad (27)$$

Further computations are conducted in accordance with the proposed procedure, using (22), (23), (24), (25) and (27) with the proper substitution of indices. Using formulas (13) and (14) we determine the coefficients of stability of the second and third stages S_2 and S_3. The coefficient of stability of the first stage is determined by the formula

$$S_1 = \frac{A_2\Delta\alpha_2 - 1}{\frac{I_{c1}\Delta\alpha_1}{\Delta I_{c3m}(\alpha_1 + \Delta\alpha_1)} - 1}, \quad (28)$$

where A_2 is determined by expression (17").

The feedback resistance R_0 can be found from the expression

$$R_0 = \frac{(S_1 - 1)(R_{L3} + R_1) - S_1 R_{L3}\alpha_1}{1 + S_1(1 - \alpha_1)}. \quad (29)$$

The magnitude of the resistance, as computed from formula (29), cannot always be acceptable, since under these circumstances the regime of the amplifier might not be satisfied.

This can be explained by the fact that the value of the resistance R_0, at a given value of U_{c3}, determines the base current of the first transistor I_{b1} and, consequently, the regime of the entire amplifier.

The expression for resistance R_0, obtained from the condition of having to satisfy the required regime of the first stage of the amplifier, equals

$$R_0 \approx \frac{(U_{c3} - U_{be1})\alpha_1 - I_{c1}R_1}{I_{c1}(1 - \alpha_1)}. \quad (30)$$

Let the magnitude of resistance R_0 as determined from formula (29) turn out to be higher than resistance R_0 as determined from formula (30). In this case, if R_0 is established, on the basis of the condition of satisfying the required regime of the first stage, then the stability of the static regime will be attained.

If, however, this condition is not satisfied, it will be necessary to repeat the calculations. Here it will be necessary either to increase the permissible value of the change in the collector current ΔI_{c3}, or to decrease the coefficients of instability of individual stages due to the increase in resistance in the transistor emitter circuits.

The AC voltage coefficient of amplification is determined as follows

$$K_u = \beta_1\beta_2\beta_3 \frac{R_{L3}}{R_{in1}} \frac{R_{L1}}{R_{L1} + R_{in2}} \frac{R_{L2}}{R_{L2} + R_{in3}}, \quad (31)$$

where

$$\beta = \alpha(1 - \alpha).$$

The value of the capacitance C is obtained from the condition of the smallness of its reactance at the amplifier operating frequency.

RESULTS OF EXPERIMENTAL CHECKS

In accordance with the proposed procedure an AC amplifier with direct feedback, using silicon transistors P103, was designed and constructed. The circuit of the amplifier is shown in Figure 2.

Amplifier specifications: E = 5 v, R_{L1} = 12 kilohms, R_{L2} = 6.8 kilohms,

R_{L3} = 3.6 kilohms, R_0 = 160 kilohms + 150 kilohms, $C_1 = C_2$ = 1 μf, C = = 10 μf. Working frequency F = 1000 cps.

Transistor parameters: at normal temperature α_1 = 0.983, α_2 = = 0.947, α_3 = 0.965, I_{C01} = 0.003 μa, I_{C02} = 0.005 μa, I_{C03} = 0.004 μa; at a temperature of +70°C α_1 = 0.989, α_2 = 0.959, α_3 = 0.973, I_C = 0.05 μa, I_{C02} = 0.2 μa, I_{C03} = 0.16 μa, ΔU_{be1} = 0.05 v, ΔU_{be2} = 0.04 v, ΔU_{be3} = = 0.04 v.

The amplifier was tested over the temperature range of +20°C to +70°C.

Calculated data: K_u = 70 db. ΔI_{C3} = 103 μa.

Experimental data: K_u = 68.8 db, ΔI_{C3} = 130 μa.

Amplitude of the output voltage U_{out} = 0.8 v.

Coefficient of amplification K_u of the amplifier during the change in temperature to +70°C increases by 1.4 db.

Computed and experimental data agree sufficiently well.

CONCLUSIONS

1. During calculation of the stability of the static regime of the amplifier, it becomes necessary to consider the effects of not only the change in the zero collector current I_{C0}, but also of the current coefficient of amplification α, for rigorous computation of the base-emitter voltage U_{be}.

2. The expressions obtained from the proposed design procedure permit the design of an AC amplifier using DC feedback, taking into account changes in α, I_{C0}, U_{be}.

3. The given formulas permit carrying out check calculations for the stability of the static operating point of a three stage amplifier with direct coupling, encompassed by negative DC feedback.

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SYNTHESIS OF SELECTIVE AMPLIFIERS WITH NEGATIVE FEEDBACK

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A method is discussed for synthesis of active RC filters, consisting of a transistorized current amplifier and a resistance — capacitance feedback network. This method permits the synthesizing of complicated RC networks, to obtain circuits with optimum number of elements and to avoid the use of multistage connection of the circuits.

The design of active RC filters, containing an amplifier with a selective resistance — capacitance negative feedback, has been a frequent topic in the literature [1, 2]. The proposed design methods basically dealt with the analysis of comparatively simple circuits possessing low selectivity. Improvement of selectivity was attained by means of a series connection of a number of simple circuits, which brings about certain difficulties when transistor amplifiers are used. The use of a synthesis method for such filters permits the attainment of RC network circuits of great complexity, the elimination of series type connections and, besides this, creates the possibility of obtaining circuits with an optimum number of elements.

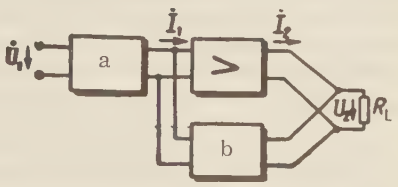

Figure 1

Below we discuss a method of synthesis of active RC filters, assembled in accordance with the circuit shown in Figure 1, consisting of a transistor current amplifier and two resistance — capacitance four-pole networks, one of which (a) is connected to the input of the amplifier and the other in the negative feedback network.

TRANSMISSION FUNCTION OF THE SYNTHESIZED CIRCUIT

Selectivity properties of the circuit of Figure 1 are completely determined by its transmission function

$$H(\omega) = \frac{\dot{U}_2}{\dot{U}_1}.$$

Complete expression for the transmission function of the circuit shown

in Figure 1, neglecting internal feedback in the amplifying element, has the form

$$H(\omega)=\frac{\alpha_T y_{21a}\left(1+\frac{y_{21b}}{g_{11T}}\right)}{(g_L+g_{22T}+y_{22b})\left[1+\frac{y_{11b}+y_{22a}}{g_{11T}}+\frac{\alpha_T y_{21b}}{g_L+g_{22T}+y_{22b}}\left(1+\frac{y_{21b}}{g_{11T}}\right)\right]}. \quad (1)$$

Here y_{21a} and y_{22a} are the elements of the $[y]$ matrix of the four-pole network a; y_{11b}, y_{21b} and y_{22b} are the corresponding terms for the four-pole network b; g_{11T} and g_{22T} are, respectively, the input and output conductances of the amplifying element when its output is short-circuited; α_T is the coefficient of current amplification of the amplifying element when its output is shorted; g_L is the conductance of the load.

Expression (1) is significantly simplified if it is assumed that the input resistance of the amplifying element is considerably smaller and the output considerably larger than the input and output resistances of the four-pole networks a and b, i.e., when the following inequalities are satisfied:

$$y_{21b} \ll g_{11T}, \quad (2)$$

$$y_{22a}+y_{11b} \ll g_{11T}. \quad (3)$$

Considering that the output conductance of the amplifying element and the conductance of the load is part of the b network, we obtain

$$H(\omega)=\frac{\alpha_T y_{21a}}{y'_{22b}\left(1+\alpha_T y_{21b}/y'_{22b}\right)}, \quad (4)$$

where $y'_{22b}=y_{22b}+g_{22T}+g_L$.

The inequalities (2) and (3) can be satisfied using the following considerations. The inequality $y_{11} \geqslant y_{21}$ holds for any passive RC four-pole network as long as $\dot{I}_2 \leqslant \dot{I}_1$. As a result of this the inequality (2) follows from (3), the one which we shall investigate

$$y_{22a}=K_1\frac{m_1(p)}{n_1(p)}; \quad y_{11b}=K_2\frac{m_2(p)}{n_2(p)}.$$

In the plane of the complex frequency p, the conductances y_{22a} and y_{11b} are determined in general form by the ratio of two polynomials.

Indicated conductances assume maximum values when $p \to \infty$. If the powers of the polynomials m_1, n_1 and m_2, n_2 are correspondingly equal then when $p \to \infty$ $y_{22a} \to K_1$ and $y_{11b} \to K_2$. Synthesis of passive RC networks is usually conducted for $K_1 = K_2 = 1$. Later on these values can be changed for the general case up to any given magnitude by changing the conductances of the circuit. By lowering K_1 and K_2, i.e., by lowering the conductances of the synthesized network, one will satisfy the inequality (3) and, therefore, the inequality (2). In the discussed case the lowering of K_1 and K_2 can not continue without a limit, since when $p = 0$ the following relation must hold

$$y'_{22b}(0) \geqslant g_{22T}+g_L; \quad (5)$$

i.e., the minimum value of the output conductance of the b network can not be less than the output conductance of the amplifying element together with the conductance of the load. In actual practice $g_{22T} \ll g_{11T}$ and in the great majority of cases one can attain the simultaneous satisfaction of inequalities (3) and (5). For this purpose it frequently becomes necessary to connect a small active resistance, which increases g_{11T}, at the input of the amplifying

element. Sometimes it is expedient to reduce the output conductance of the amplifying element by the introduction of substantial negative feedback. Both methods, however, lead to a reduction of the coefficient of current amplification α_T and to an increase in the number of amplifier stages.

If inequalities (2) and (3) are satisfied it will not only simplify the process of network synthesis, but also decrease the effect of changes in the output and input conductances of the amplifying element on the frequency characteristic of amplification of the synthesized circuit. The effect of change in the current amplification coefficient α_T on this characteristic will be discussed below.

DETERMINATION OF PARAMETERS OF RC NETWORKS a AND b

In the synthesis process we assume that the function which describes the behavior of the synthesized circuit in the complex frequency plane is given. In the general case this function is given in the form of a ratio of two polynomials.

$$H(p) = A\frac{(p-p_1)(p-p_3)(p-p_5)\dots}{(p-p_2)(p-p_4)(p-p_6)\dots} = A\frac{M(p)}{N(p)}. \tag{6}$$

The poles of the function p_2, p_4, etc., as a rule, are complex and, consequently, it is impossible to realize the given function by means of a passive RC network [3].

The basic problem of the presented method of analysis consists of the determination of the parameters of the four-pole networks a and b, i. e., the magnitudes y_{21a}, y_{21b}, y_{22b} and α_T.

Let us present the polynomial N(p) in the form

$$N(p) = B_1N_1(p) + B_2N_2(p), \tag{7}$$

where B_1 and B_2 are substantial positive numbers. Considering (7), expression (6) may be rewritten

$$H(p) = A\frac{M(p)}{B_1N_1(p)\left[1+\frac{B_2}{B_1}\frac{N_2(p)}{N_1(p)}\right]}. \tag{8}$$

Let us select some polynomial with substantial negative roots Q(p) and let us write (8) in the form

$$H(p) = \frac{A\cdot M(p)/Q(p)}{B_1\cdot N_1(p)/Q(p)\left[1+\frac{B_2}{B_1}\frac{N_2(p)/Q(p)}{N_1(p)/Q(p)}\right]}. \tag{9}$$

Comparing (9) with (4) we obtain

$$y_{21a} = M(p)/Q(p), \tag{10}$$

$$y'_{22b} = y_{22b} + g_{22T} + g_L = N_1(p)/Q(p), \tag{11}$$

$$y_{21b} = N_2(p)/Q(p), \tag{12}$$

$$\alpha_T = B_2/B_1. \tag{13}$$

For physical realization of the four-pole networks a and b in the form of resistance — capacitance networks, it is necessary to set certain limitations for the polynomials $N_1(p)$, $N_2(p)$ and Q(p): 1) the roots of the polynomial $N_1(p)$ must be real and negative; 2) the roots of polynomial $N_1(p)$ and Q(p) must alternate; the root with the smallest modulus must belong to

the polynomial $N_1(p)$; 3) the power of the polynomial $Q(p)$ must equal the power of the polynomial $N_1(p)$ in accordance with the considerations presented above; 4) polynomial $N_1(p)$ may not possess a root $p_i = 0$, otherwise it will be impossible to extract the active conductance $g_{22T} + g_L$ from the four-pole network b, as it follows from (11).

Aside from the statements made above, it is desirable to limit the roots of the polynomial $N_2(p)$ to the left semiplane; with this the four-pole network b becomes a network of a minimal phase type. This situation makes it easier to maintain the stability of the synthesized amplifier outside the operating frequency band. The selection of the polynomial $N_1(p)$, taking into account the indicated limitations, is to a certain extent arbitrary. However, the instability of the frequency characteristic of the synthesized filter, caused by the instability of the coefficient of current amplification of the amplifying element, depends to a great extent upon the selection of the roots of this polynomial. Actually the instability of the frequency characteristic of the synthesized circuit can be evaluated by the ratio of the percentage change in the modulus of the transmission function $H(i\omega)$ to the percentage change of the amplifying element's coefficient of current amplification. This ratio may be called the sensitivity of the circuit

$$S = \frac{d|H(i\omega)|}{|H(i\omega)|} \Big/ \frac{d\alpha_T}{\alpha_T}. \tag{14}$$

From (4), taking into account (10)-(13), one easily obtains

$$S = \frac{1 + \alpha_T |N_2(i\omega)/N_1(i\omega)| \cos[\varphi_2(\omega) - \varphi_1(\omega)]}{1 + 2\alpha_T |N_2(i\omega)/N_1(i\omega)| \cos[\varphi_2(\omega) - \varphi_1(\omega)] + \alpha_T^2 |N_2(i\omega)/N_1(i\omega)|^2}, \tag{15}$$

where $\varphi_1(\omega)$ and $\varphi_2(\omega)$ are the arguments of the polynomials $N_1(i\omega)$ and $N_2(i\omega)$, respectively. A consideration of the ratios (7) and (13) permits (15) to be brought to the following form

$$S = B_1 |N_1(i\omega)/N(i\omega)| \cos[\varphi_1(\omega) - \varphi(\omega)], \tag{16}$$

where $\varphi(\omega)$ is the argument of the polynomial $N(i\omega)$.

The last expression shows that the sensitivity of the circuit depends upon the selection of the roots of the polynomial $N_1(p)$. In particular these roots must be selected in the vicinity of the complex frequency plane coordinate origin, since in that case the value $|N_1(i\omega)|$ will be a minimum. In the selection of the roots of the indicated polynomial one should also consider the changing of the arguments $\varphi_1(\omega)$ and $\varphi(\omega)$ such that their difference be a maximum possible and be close to $\pi/2$. Using the selected polynomial $N_1(p)$ one can determine the roots of the polynomial $N_2(p)$ from (7)

$$N(p) - B_1 N_1(p) = 0. \tag{17}$$

If the power of the given function $H(p)$ is sufficiently high, difficulties will arise when Equation (17) is solved. One way of surmounting these difficulties is to make use of the method of root hodographs [4]. Expression (17) may be rewritten as

$$1 - B_1 N_1(p)/N(p) = 0. \tag{18}$$

From here

$$|N_1(p)/N(p)| = \frac{1}{B_1}; \quad \arg[N_1(p)/N(p)] = 2k\pi. \tag{19}$$

In this manner the roots of the polynomial lie on hodographs determined by the equalities (19). To each point on the hodograph there

corresponds a concrete value of B_1. The selection of B_1 is determined by the requirements stated above applied to the polynomial $N_2(p)$. Aside from that, in the selection of B_1 one should consider its effect on the sensitivity of the circuit (16) and also upon the value of the amplifying element current amplification coefficient (13). From the above it follows that the value of B_1 should be selected to be as small as possible, however, so that the obtained current coefficient of amplification α_T must not turn out to be unduly high. When the powers of the polynomials $N_1(p)$ and $N_2(p)$ are equal the following relation will hold:

$$B_1 + B_2 = 1. \tag{20}$$

For this we obtain respectively for the coefficient of current application α_T and the sensitivity S

$$\alpha_T = \frac{1 - B_1}{B_1}. \tag{21}$$

Polynomial Q(p) is selected in accordance with the considerations given above:

$$S = \frac{1}{1 + \alpha_T} |N_1(i\omega)/N(i\omega)| \cos[\varphi_1(\omega) - \varphi(\omega)]. \tag{22}$$

The presented method of synthesis permits the determination of the parameters of networks a and b of the circuit of Figure 1. Using these parameters one can synthesize, by known methods, the resistance — capacitance four-pole networks [4]. The most acceptable method in this case is the Dascher method.

The majority of existing methods of RC network synthesis permit one to synthesize the circuits for a given transmission function within the accuracy of a constant multiplier. Thus, for the realized feedback of the four-pole network we may write

$$\frac{y_{21\,b}}{y_{22\,b}} = K \frac{N_2(p)}{N_1(p)}. \tag{23}$$

Hence, it is evident that the coefficient of current amplification of the realized amplifying element α_T^* must differ from the computed value, as obtained from (21), by a factor of K

$$\alpha_T^* = \alpha_T / K. \tag{24}$$

STABILITY OF THE SYNTHESIZED CIRCUIT

In the above discussion it was stated that the current amplification factor of the element is a substantial quantity and does not depend on frequency. At relatively low frequencies and for the condition where the amplifying element does not contain dividing capacitors this assumption is entirely justified. At these frequencies the synthesized circuit will be stable as long as the zeros of the polynomial $N_2(p)$ are limited to the left semiplane of the complex frequency. At higher frequencies, however, the coefficient of amplification turns out to be complex and dependent upon frequency. The phase shift introduced by the amplifying element at these frequencies may lead to oscillation of the entire amplifier circuit, if at the frequency at which the phase shift along the feedback loop equals $2k\pi$, the current amplification factor in the loop does not drop to a value smaller than unity.

At sufficiently high frequencies the coefficient of current transmission in the feedback network can be considered to be equal to K. Actually, as long as the powers of the polynomials $N_1(p)$ and $R_2(p)$ are equal, it can be seen from (23) that for a sufficiently high frequency

$$\left[K\frac{N_2(p)}{N_1(p)}\right]_{p\to\infty}\to K. \tag{25}$$

Hence, the coefficient of current amplification along the feedback loop, taking (24) into account, at high frequencies is equal to

$$\alpha_T^{\bullet}K=\alpha_T. \tag{26}$$

The amplifying stage of the transistor, for common emitter, introduces a current phase shift of the order of 45° at frequencies approaching the critical frequency of the transistor. Therefore, for example, we may expect the appearance of a three-stage amplifier at a frequency somewhat higher than the critical. In order that a required reservoir of stability be attained it becomes necessary to introduce into the circuit of the amplifying element, compensating elements, in accordance with the procedure presented in the known monograph of Bode [3]. If the synthesized circuit must guarantee a given frequency attenuation characteristic within the boundaries not outside the limits of 0.1 of the critical frequency, then the introduction of the compensating elements will practically not bring about any noticeable effects on this characteristic.

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ANALOG SIMULATION OF SWITCHING CIRCUITS

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The report discusses a method of determining the capacity of switching circuits by static trials on universal and special-purpose electronic computers. Examples are given for analog machines for solution of some specific problems.

Problems dealing with the capacity of switching circuits are of great importance in the theory of telephone communications. This subject has recently attracted great attention in connection with the development of the general theory of information distribution. In order to avoid misunderstanding we will note that the term "capacity" is used in two senses. In information theory this term is used for the characteristic of a communications channel (or a communications system as a combination of technical facilities for the transmission of information from sender to receiver). In this case the capacity is defined as the amount of information which may be transmitted by a communications channel per unit of time. In telephone communications theory this term is used for the characteristic of a network consisting of switching devices and the lines between them and serving to connect subscribers. Here the capacity is defined as the number of connections (or channels) which may be created between different pairs of subscribers per unit of time equal to the average duration of one connection. In other words, in this case capacity is evaluated in terms of the traffic. In this report we shall refer to capacity in the second sense of the term.

Problems of capacity of a switching circuit are solved by the methods of the mathematical theory of mass service. For a wide class of switching circuits characterized by the fact that the incoming flow is the simplest, incompleted calls are lost and the holding time has an exponential distribution. This theory describes the operation of the circuit as a unidimensional or multidimensional Markoff process [1, 2]. Determination of the capacity in this case reduces to calculation of the loss probability as the sum of final probabilities of states leading to incompleted calls. It must be noted that in theoretical investigations in recent years the theory of Markoff processes is used also with more general assumptions (for an arbitrary flow, arbitrary distribution of holding time, for systems with delay, etc.). However, such investigations are limited to only the simplest structure, a completely accessible trunk line [3, 4]. For systems of mass service used in communications a characteristic feature is complexity of structure [5, 6]. Hence, the specific difficulties in solving the problem due to the fact that the circuit may have many different states and determining the final probabilities of the states of interest to us reduces to obtaining and solving systems of equations with an extremely large number of unknowns [7]. For practical circuits the execution of these operations is impossible even on

electronic computers. Hence, the problem of capacity of switching circuits is most effectively solved by statistical trial methods (which in the foreign literature is known as the "Monte Carlo method"), that is, by analog simulation [8]. Simulation is useful not only for determining the capacity of circuits whose operation is susceptible to theoretical description but to the solution of problems confronted with insurmountable practical difficulties. At the present time the theory of mass operation still does not provide an answer to all practical questions. This necessitates the use of engineering design methods which are sometimes without adequate theoretical foundation. In view of this the simulation method is useful both in determining the capacity of circuits whose operation is not described in theory and for checking applied engineering design methods.

PRINCIPLE OF STATISTICAL TRIALS

In the general case the investigation of capacity of a circuit by the statistical trial method is performed with the following block diagram (Figure 1). Here 1 is a device simulating the process of arrival of calls; 2 is the analog of the switching system under test or, in general, any realization of the algorithm according to which the investigated system operates with a given distribution of duration of connections; 3 represents recording devices. The properties of the simulated circuit which are of interest to us are determined as the result of trials of the model over a given period of time and by subsequent statistical processing of the resulting data. Investigation with the arrangement shown in this block diagram may be achieved both on a general-purpose digital computer (GPDC) and by means of a special-purpose analog machine.

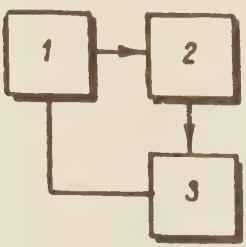

Figure 1

In Figure 1 the first section is the source of load (for example, a device simulating the behavior of telephone subscribers). This device may be a random-number transducer. Let us first demonstrate this for the simplest case. From the theory of mass operation [1] we know two equivalent methods of giving the flow of calls as a random process. On the one hand, the flow may be given by expressing as function $f(t_i)$ the distribution of the number of calls arriving in each interval of time t_i. On the other hand, in a completely analogous manner we may give the distribution of duration of intervals between adjacent calls $Z_i = T_i - T_{i-1}$, where T_i and T_{i-1} are the instants of access of the i-th and (i-1)th calls. Flows with mutually independent intervals Z_i are known as flows with limited consequence. For such flows the distribution of duration of the first interval may be written in the form

$$F_i(t) = \lambda \int_0^t \varphi_0(u)\, du, \tag{1}$$

and for the i-th interval ($i > 2$)

$$F_i(t) = 1 - \varphi_0(t). \tag{2}$$

In these expressions λ is the intensity of flow and $\varphi_0(t)$ is the Palm function for the probability of absence of calls in the interval t on the condition that at the initial moment of this interval a call has arrived. For the

simplest flow the conditional probability $\varphi_0(t)$ is equal to the unconditional probability $v_0(t) = e^{-\lambda t}$ and, consequently,

$$F_1(t) = F_i(t) = 1 - e^{-\lambda t}. \quad (3)$$

Expressions (1-3) are the basis of simulation of flows by means of random-number transducers. But in order for a digital computer to operate with a continuous function, the time is quantized and represented as a dimensionless quantity (the number of machine operations). Then the probability that a call will not arrive during k operations is $v_0(k) = (1-p)^k$ and the distribution of interval duration to the next call has the form: $F_i(k) = 1-(1-p)^k$. In these expressions p is the probability of arrival of a call during one operation Δt. In the limiting case

$$v_0(t) = \lim_{\Delta t \to 0} (1 - \lambda \Delta t)^{t/\Delta t} = e^{-\lambda t} \text{ and } F_i(t) = 1 - e^{-\lambda t}.$$

Thus, the exponential distribution described in formula (3) is replaced by geometric distribution with coefficient 1-p = q. The random-number transducer designed for simulation of such a flow must deliver a series of numbers with geometric distribution. With small Δt

$$p = 1 - e^{-\lambda \Delta t} \approx \lambda \Delta t,$$

that is, the parameter of the simulated flow is proportional to the probability of occurrence of a given number (or group of numbers) in the transducer. The probability of arrival of i calls during k operations follows a Bernoulli distribution

$$v_i(k) = \binom{k}{i} p^i (1-p)^{k-i}.$$

But, as is known, within the limit, when $\Delta t \to 0$, $v_i(t) \to e^{-\lambda t} \frac{(\lambda t)^i}{i!}$, that is, the random-number transducer provides a Poisson process with a certain accuracy.

The model of the switching system to be tested is often built along the lines of an analog computer. However, there arise difficulties in achieving a given distribution of duration of connections and the system of operation. It is simplest to make the analog of a switching system with incompleted calls and exponential distribution of holding time. In this case Section 1 (Figure 1) controls clearing. In effect, the probability that a given hold will last not less than t units of time is $e^{-\beta t}$, where $1/\beta = t$ is the average holding time. This principle is simulated by the same random-number transducer as controls the arrival of calls. The same principle is employed in simulating a system with delayed ringing [9].

SIMPLEST FLOW AND EXPONENTIAL DISTRIBUTION OF HOLDING TIME

On the basis of the presented principles we may formulate the chief requirements for a random-number transducer designed for simulation of the simplest flow with exponential distribution of holding time. The transducer must represent (1, M) - pole, the input of which is periodically acted upon by clock pulses while at one of the outputs during each cycle in a random sequence there appears a response in the form of an output pulse [10]. The value of the random number is determined by the number of the output at which the pulse appears. (Such a random-number transducer is sometimes referred to as a "ruletka".) The probability of occurrence of

a given number during each operation of the transducer is p = 1/M, where M is the total number of transducer outputs (i.e., the number of possible single-digit numbers or the radix of the system). The assembly average of the number of operations before the next occurrence of the designated single-digit number is

$$E(x) = \sum_{i=0}^{\infty} (1 + i)\, p\, (1 - p)^i = \frac{1}{p} = M.$$

If the average holding time $1/\beta$ is also made equal to M (this depends on the chosen time scale), then clearing of each line must be simulated by the occurrence of one of the numbers. In order to simulate the flow of load in λ Ehrlangs, λ numbers are required, since the load is defined as the assembly average of the number of calls per average time of one hold. If it is required to construct an analog with the possibility of load control from 0 to A Ehrlangs with steps of a Ehrlangs in g load groups with the total possible number of lines n (the arrangement for connection of these lines is of no importance and may vary), then we must provide for

$$M = n + Ag \text{ outputs, if a is an integer} \tag{4}$$

$$\text{or } M = \frac{n + Ag}{a} \text{ outputs, if } a = \frac{1}{b} \text{ is a simple fraction} \tag{5}$$

Here M is the smallest possible radix (or the minimum number of code combinations in the case of use of a general-purpose digital computer). An increase in number M provides closer approximation to a Poisson flow (since there is a decrease in p = 1/M), but is accompanied by an increase in machine time required for the experiment, since there is a decrease in the time quantization level. Hence M must be chosen with consideration of the lower limit defined by formulas (4) or (5) keeping in mind the permissible error in approximating the law of distribution. This error is determined by the time scale adopted for the analog computation. If the average holding time in the simulated system is T seconds and is simulated in the machine by means of N operations, the time quantization level will be $\Delta t = T/N$ seconds and the analog computation error may be evaluated by the quantity $\epsilon = \frac{1}{2}\frac{\Delta t}{T}\,100\% = \frac{100}{2N}\,\%$. In special-purpose machines designed for investigation of switching circuits random-number transducers with M outputs are used, wherein those inputs in excess of the number defined by the righthand member of expression (4) or (5) may not be activated. The code combinations of binary random numbers in the general purpose digital computer must be chosen with that number of digits m which satisfies the inequality $2^m \geq M$.

The possibility of using uniformly distributed random numbers for simulation of simplest flow with exponential distribution in servicing was known in telephony long before the appearance of electronic computers. There was a time when wide use was made of a method of investigating switching circuits by analog computation "on paper" by means of tables of random numbers and recording of the states of the system (the so-called "artifical telephone communication" method). Special-purpose analog machines were later used for such investigations. The wide application of electronic computers now permits such analog computation to be performed on general-purpose digital computers. On the other hand, due to the complexity of programming and large machine time requirements in solving

problems by statistical trials with general-purpose digital computers, the use of special-purpose computers is under consideration for this purpose.

As an example of such a computer let us take an analog machine for investigating incomplete-access circuits based on standard three-stable-state ferrite cells (Figure 2). For simplification this figure shows the analog of the simplest circuit with n = 11 outputs, number of loading groups g = 6 and accessibility D = 3 (Figure 2a). The corresponding analog device is represented in Figure 2b, showing a random-pulse transducer with M outputs and an analog of the investigated circuit. The recording devices indicating time losses and call losses are not shown in the figure. Cell A represents an "AND" gate and cell T is a dynamic flip-flop which is triggered upon entrance of a pulse at the upper input and disconnected upon entrance of a pulse at the lower input. The analog of a separate element of the circuit (output) is a device having two stable states and operating as follows (Figure 2c). If a pulse reaches input 2, the device is in its first state (the dynamic flip-flop is out of the circuit and the "line" is clear). If a single pulse then reaches input 1, the device switches to its second state (the dynamic flip-flop is triggered and the "line" is busy). In this state all subsequent pulses reaching input 1 will pass to the output. This state will exist until at least one pulse reaches input 2, which pulse will return the circuit to the first state (the "line" is cleared). In other words, between input 1 and the output there acts a special type of valve opening along input 1 and closing along input 2. For the sake of brevity we shall refer to the

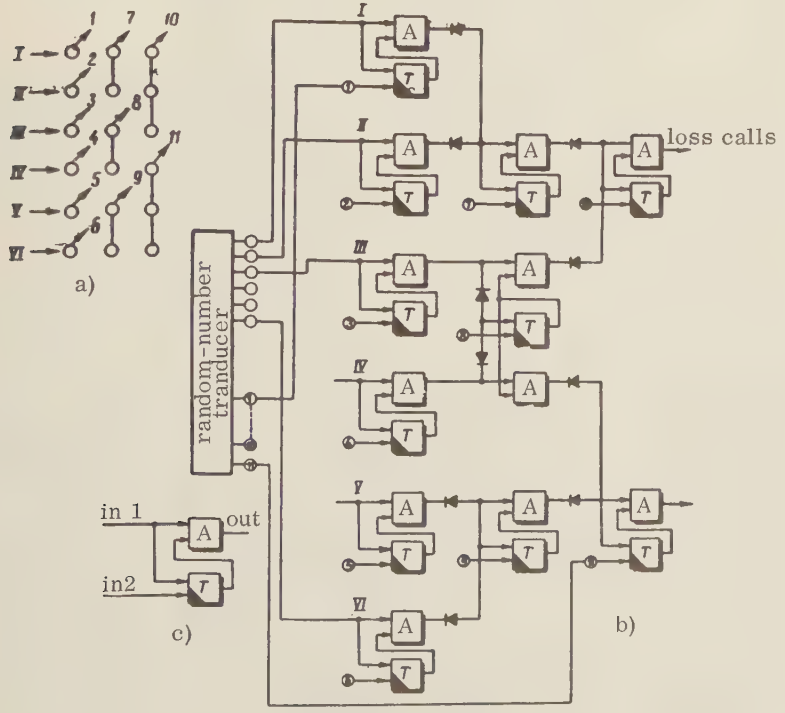

Figure 2

described device as a pulse valve. With buffer diodes such pulse valves may be used to construct any incomplete-access switching circuit, including more complex circuits of practical interest. A distinctive feature of the connection of elements is the fact that the T cells (simulating common lines whose contacts have further access to different lines — the eighth line in Figure 2) are joined with several A cells, forming, as it were, a multichannel valve. In designing a model of the described type it is necessary to consider the time required for information to travel through the cells, which necessitates a reduction in operating speed of the analog device. This disadvantage is encountered in transistorized analogs, the design principles of which are quite similar.

As another example of investigation of capacity in the case of simplest flow and exponential distribution of holding time Figure 3a shows the block diagram of a routine for checking the group selector of a crossbar exchange as programmed in Sweden for a general-purpose digital computer, the BESK [11]. This example is of particular interest as an attempt to use a general-purpose digital computer for solution of problems in the theory of

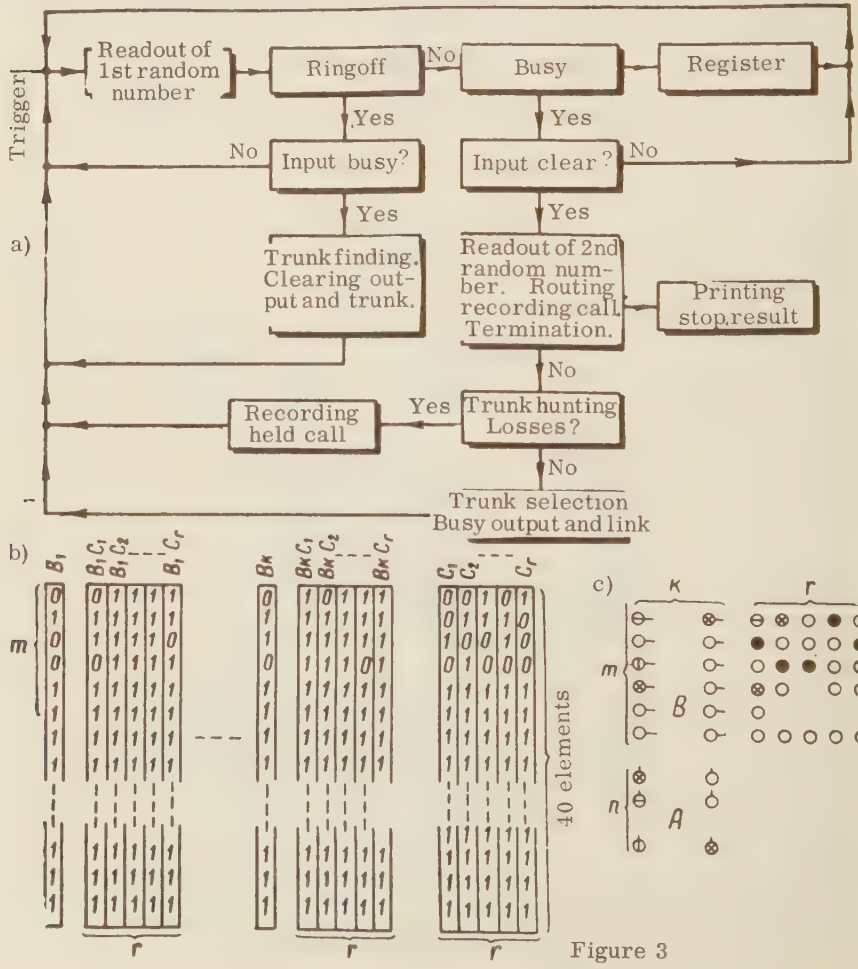

Figure 3

telephone communication. In principle, a similar analog may be constructed by means of pulse valves, but in this case a vastly greater number of such elements would be required than in the circuit in Figure 2. An idea of the required size of the memory is given in Figure 3b, showing the diagram of the memory of the general-purpose digital computer. Here the ones and zeroes represent cleared and busy cells. The corresponding two-section circuit with notations used in crossbar exchanges is shown in Figure 3c. This notation permits us to ascertain what connections exist in the circuit in the discussed state. The checking routine for the general-purpose digital computer is so compiled that the tests are performed in a manner similar to the analog in Figure 2. The only difference lies in the fact that in the BESK computer pseudorandom numbers are obtained. In addition, the analog of the two-section circuit requires a large number of analog elements; hence the memory of the general-purpose digital computer is quite convenient for storage of information concerning the state of large circuits. But the pseudorandom numbers obtained by the program method (that is, a deterministic method) are not random in the probability-theory sense of this term. Hence, even with general-purpose digital computers the use of special random-number transducers is preferred, insuring substantial reliability in investigation. Statistical trials on a general-purpose digital computer require considerable machine time. In this connection there is increasing adherence to the opinion that it is necessary to specialize digital computers for the purpose of reducing the time expended in statistical trials.

ANALOG COMPUTATION IN THE MOST GENERAL CASES

It is well known that the case of simplest flow and exponential distribution of call duration is relatively simple for investigation and extremely important in telephone practice. However, in addition to this case, there are many other practical problems to be solved, the theoretical investigation of which is attended by considerable difficulties. Hence, the analog computation of such problems is of considerable interest. We shall dwell at length on the specific problems of analog computation by the use of general-purpose digital computers in view of the fact that there is as yet little experience in this field. There will undoubtedly be a considerable accumulation of experience in future years and it will be possible to obtain appropriate conclusions. Hence, let us discuss several examples of analog computation on special-purpose computers. These examples may also be applied in the use of general-purpose digital computers.

The random-transducer is, of course, the first element required in the analog computation. Such a transducer provides a Poisson flow with sufficient accuracy. However, with certain additions such a transducer may also provide several classes of unsteady flow and also flows with limited secondary effects which are of interest in telephony. For example, with a limited number of load sources the call traffic will not be a Poisson flow, since its intensity decreases in proportion to the number of busy lines. The mathematical description of the flow in this case is known as an Engset function. In the analog computation of such a flow it is necessary that there be provisions within the computer for isolating the random-number transducer output from the load busses in the presence of busy lines. This is achieved by means of so-called artifical subscribers [6]. Schematically these are suggestive of pulse valves. They also have busy inputs and cleared inputs. However, in distinction from a pulse valve, upon

the access of a pulse to a busy input the artificial subscriber is itself busy and transmits the busy condition to the analog. In this state the subsequent pulses reaching the busy input are lost. Upon the access of a pulse to the clearing input the artificial subscriber is cleared and the corresponding line of the analog is cleared.

As another example let us take the analog computation of several classes of flows with limited after effect. If the Poisson flow is filtered by means of a network of pulse valves PV (Figure 4), at the output of each PV there is obtained the corresponding flow with limited aftereffect, the

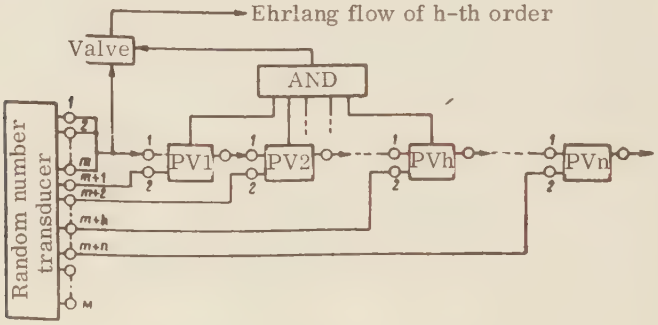

Figure 4

output of each PV there is obtained the corresponding flow with limited after effect, the intensity of which is defined by the Ehrlang formula. If, in addition, in this circuit we provide for a check of the simultaneous busy state of the first h PV (for example, by means of an "AND" gate), we may obtain the so-called Ehrlang flows. These are characterized by the fact that the number of calls incoming during a given time segment t obeys an Ehrlang distribution function

$$v_k(t) = \frac{\frac{(\lambda t)^k}{k!}}{\sum_{i=0}^{n} \frac{(\lambda t)^i}{i!}},$$

where h is the order of the given Ehrlang flow. For this purpose it is necessary to pass a Poisson flow through the valve controlled by the "AND" gate. The order h of the Ehrlang flow is determined by the number of pulse valves connected with the "AND" gate. The pulse valves may use various electromagnetic or electronic devices, depending on the required operating speed. In the analog computation of systems with delay, particular difficulties arise due to the fact that in such systems there is inadequate measurement of losses and it is necessary to determine the delay-time distribution. In this case it is particularly necessary to conform to the time scale, which is determined on the basis of the following considerations. Assuming a required quantization error ε, the quantization level Δt and number of operations N corresponding to average call duration (as well as the number of operations corresponding to a single hour of maximum load) are determined.

Upon determining the time scale and assigning the permissible statistical error ε_c, the required volume of statistical trials [8] is determined from the formula $\varepsilon_c = 3\sigma/\sqrt{K}$, where K is the volume of statistical trials per period (in time units of T) and σ is the variance of the random quantity to be determined. If the variance is not known beforehand, we may employ empirical evaluation of the variance as determined in the process of testing. After the volume of trials has been determined it is not difficult to find the total time expended in experiment

$$T_{exp} = \frac{9\sigma^2}{\varepsilon_c^2} v t_0,$$

where v is the average number of operations per trial and t_0 is the time expended per operation.

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START-STOP RECEIVING DEVICES OPERATING WITHOUT INTERRUPTION OF CONTROL PULSE GENERATOR

P. A. Kotov

The report discusses two circuits for a start-stop receiving device using ferrite cores with rectangular hysteresis loop (RHL) and transistors operating without interruption of the control pulse generator. The circuits are distinguished by high operating characteristics.

PURPOSE OF DEVICES

In telegraph systems with automatic detection of error and interrogation (AEI) for reception of information from correspondents, standard start-stop reperforators are used.

However, in those cases when the AEI system is connected for operation along wideband channels which permit a high rate of transmission, the use of an ordinary reperforator for reception of information from the terminal station is not possible. In the given case, for reception, it is necessary to have high-speed start-stop receiving equipment, and storage of the received information is achieved on magnetic tape or other high-speed devices.

The literature contains much material describing high-speed start-stop receiving devices using vacuum tubes, transistors and other contactless elements. However, the better known circuits employ an individual control pulse generator (CPG) the oscillations of which are interrupted during the stop period. Due to instability of the oscillatory phase of the generator upon triggering, start-stop receiving devices built on this principle have considerable phase error.

Simplicity and operational stability are characteristic of the high-speed start-stop receivers with contactless switches without interruption of control pulse generator and are discussed below. These circuits may also be used in constructing high-speed start-stop teleprinters, start-stop regenerators and reperforators comprising start-stop synchronous telegraph systems (STS).

Start-stop receivers, which do not use interruption of the control pulse generator and which do use contactless switches, have the following properties.

1. Stop-holding of the start-stop receiver is achieved without interruption of the control pulse generator. This permits the use of a common,

highly stable generator for shaping the control pulses for the entire communications link. Moreover, by the use of start-stop receivers which do not operate by interrupting the control-pulse generator and by the use of devices for phase correction of synchronous distributors which do not affect generator parameters, all of the required conditions are created for development of a new communications system for transmission of discrete information in which control of all the transmitting, receiving and relaying devices at each station is achieved by one highly stable generator. In this case phase correction of the control pulses may be centralized by transmission of correcting pulses, not over all the individual communications channels, but along only one highly stable channel. A communications system with centralized correction considerably simplifies the operational servicing of terminal equipment and increases the stability and reliability of its operation.

2. The start-stop receivers are started only when the duration of the start pulse is not less than $t_0/2$, where t_0 is the elementary pulse duration. Thus, the discussed start-stop receivers are protected against interruption during stops by short pulse interference.

3. Interruption of a start-stop recievers at the stop position is ensured in case of considerable distortion of the stop pulse: at the sending end the distortion may be reduced by approximately 90-95% and at the receiving end it may be reduced by 48-49% without disturbing normal operation of the receiver.

START-STOP TRANSISTOR RECEIVERS USING FERRITE CORES WITH RECTANGULAR HYSTERESIS LOOP

The diagram of a transistorized start-stop receiver which does not interrupt the control pulse is shown in Figure 1. The circuit consists of a receiving relay RR, a flip-flop T_6 and T_7 and two "AND" gates.

The circuit operates as follows. In the "stop" state the control flip-flops T_6 and T_7 and all flip-flops of the frequency dividers are in the state in which high voltage exists at their unshaded areas B. This state of the flip-flops is ensured by operation of the corresponding elements of the

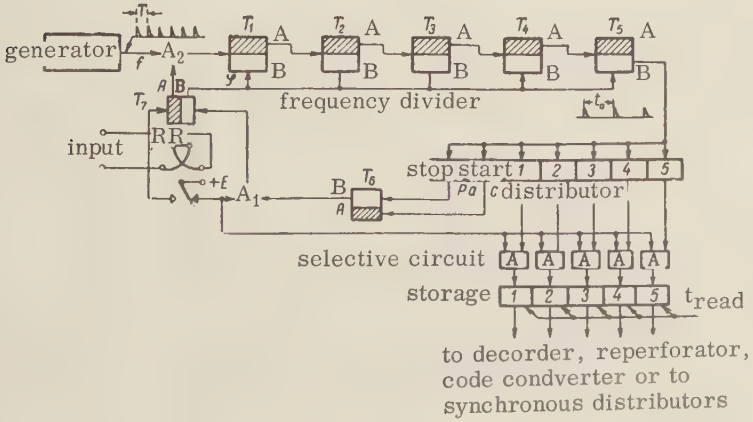

Figure 1

circuit, which is easily seen from examination of the operating principle of the circuit. During reception of the stop signals from the "stop" distributor a pulse reaches the input of flip-flop T_6, under the action of which it is reversed and from its B output there is applied a high voltage to AND_2. During reception of the stop signal the RR armature is located at the right hand contact, with the result that there will also be applied a high voltage to the second input of AND_1. From the output of AND_1 high voltage reaches the input B of flip-flop T_7, which reverses and gives a high voltage at output B, and a low voltage at output A. If from output A of flip-flop T_7 high voltage is not obtained, the pulses from the generator will not pass through AND_2 and pulse φ from output B of flip-flop T_7 sets all flip-flops of the frequency divider to the state indicated in the figure, in which state high voltage will exist at their B outputs. In this state the circuit will be in a condition to obtain the start pulse.

Upon receiving the start pulse the RR armature is first displaced from the right-hand contact, with the result that application of high voltage to input B of flip-flop T_7 is interrupted. Then upon closing the RR armature, by means of the left-hand contact a pulse will be applied to input A of flip-flop T_7 and the flip-flop will switch to another state in which state its output A applies a high voltage to AND_2. From this instant the control pulses from the generator can pass through AND_2 to the input of the frequency divider.

With the first control pulse all flip-flops of the frequency divider will transfer to the state in which outputs A are at high voltage. At this instant operation of the start-stop receiver begins.

At the time of receiving the middle portion of the start pulse from output B of flip-flop T_5, a pulse will emanate from the frequency divider, converting the distributor to its next stable state, and at this instant from its "start" contact a short pulse is transmitted to input A of flip-flop T_6. Flip-flop T_6 will change to its other stable state, where its output B is at low voltage and AND_2 will be blocked for the entire period of reception of the code combination. Consequently, from output A of flip-flop T_7 during the entire period of reception of the code combination at the input of AND_2 there will be applied a high voltage and the control pulses from the generator will flow unblocked to the input of the frequency divider. At the end of the cycle, upon reception of the stop pulse from the stop" contact of the distributor, there will appear a pulse at B of flip-flop T_6 which will switch it to its other stable state and to the input of AND_1 a high voltage will be applied. As a result of this the receiver will change to the "stop" state shown in Figure 1.

It is easily seen that upon the access of a start pulse with duration less than $t_0/2$ at the input of the receiver the armature of receiving relay RR will switch to the right-hand contact until the instant of arrival of the pulse from output B of flip-flop T_5 of the frequency divider. As a result of this, there will pass through AND_1 a pulse which will reverse flip-flop T_7 to its other stable state, in which high voltage exists at output B. Application of control pulses to the frequency divider input will be interrupted and its flip-flops will be set to the initial position by pulse φ.
It is also easily seen that the discussed start-stop receiver will be safely set to the "stop" state with an initial pulse distortion of 90-95%.

The discussed start-stop receiver circuit is suitable for start-stop teleprinters, reperforators and associated equipment of start-stop telegraph systems. It must be pointed out that in Figure 1 the receiving relay is represented as an electromagnetic relay for purposes of simplification of the diagram.

Figure 2 shows the diagram of receiving and transmitting devices

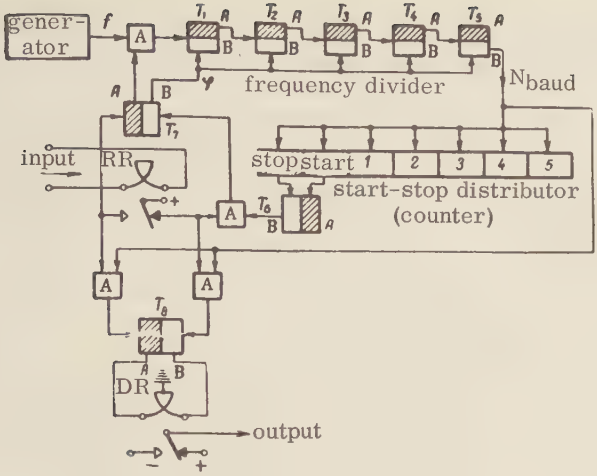

Figure 2

constructed for a start-stop regenerator. This diagram differs from the first diagram in that the distributor serves as a cycle-rate counter.

Figure 3 shows the operational diagram of the frequency divider of a start-stop receiver. It is seen from the diagram that the entire frequency divider cycle is completed during reception of one elementary pulse of duration t_0. At the moment of reversal of flip-flop T_5 of the frequency divider from state A to state B, from one of the distributor contacts there is transmitted a pulse which is used as a selector pulse. It is also seen from the diagram that at the moment of reception of the middle portion of the stop pulse, all flip-flops of the frequency divider are set by pulse φ to the initial (stop) state (Figures 1 and 2).

Figure 4a shows another version of a start-stop receiver in which the frequency divider, distributor and control elements use ferrite cores with rectangular hysteresis loop. In this circuit, in the "stop" state the device prohibiting application of control pulses to the input of frequency divider D_1 is a dynamic flip-flop DT which is triggered by pulse P from output M_2 upon reception of the start signal and is stopped by pulse 0 from output M_1 upon reception of the stop signal.

The circuit operates as follows. In the "stop" state dynamic flip-flop

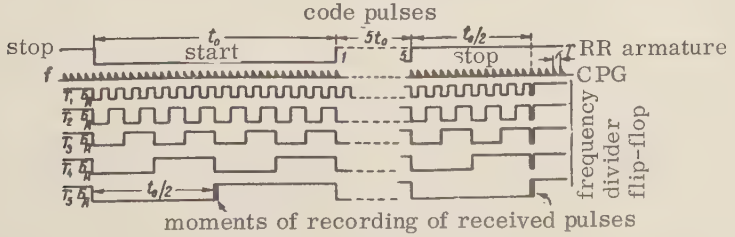

Figure 3

DT does not operate and no control pulses reach the input of frequency divider D_1. Under the action of a pulse from the "stop" contact of the distributor, flip-flop T is placed in the state at which high voltage is at output B.

Upon reception of the start pulse, the armature of RR switches to the left-hand contact and in core M_2 through AND_2 there will be stored 1. Upon the arrival of sequence pulse e_2 from the output of frequency divider D_0 the stored 1 is read out from core M_2 and 1 is stored into core DT. From this instant from the output of DT control pulses will arrive at the input of frequency divider D_1.

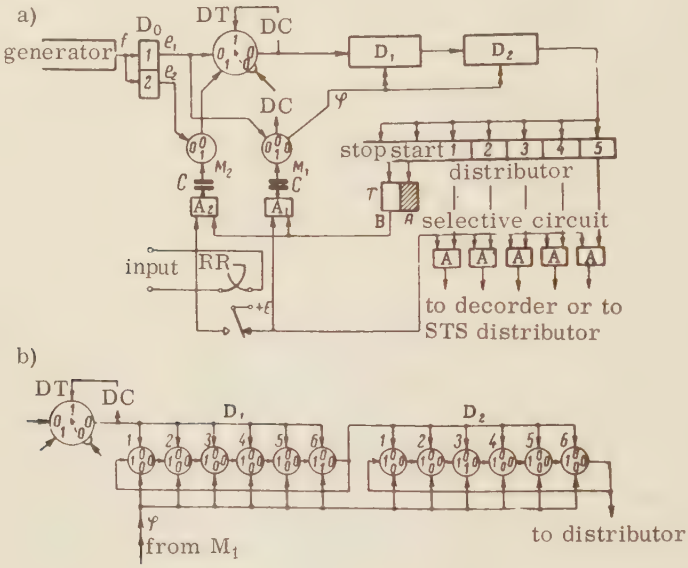

Figure 4

Upon the reception of the middle portion of the start pulse from the output of the frequency divider there will arrive a pulse which changes the distributor to a new stable state and from its "start" contact there arrives a pulse which sets flip-flop T to the state at which high voltage exists at output A. From this moment cells AND_1 and AND_2 will be closed for the entire cycle of reception of the code combination.

At the end of the cycle and at the moment of reception of the middle portion of the start signal from the "stop" contact of the distributor there will arrive a pulse which will switch flip-flop T to its other state and from its output there will be applied to AND_1 and AND_2 a high voltage. If at this time the armature of RR is at the right-hand contact, through AND_1 there will be entered in core M_1 the digit 1. Upon readout of this 1 by pulse e_1 from the output of core M_1 through the delay circuit DC to the input of DT there arrives a pulse which blocks the recording of 1 by the pulse traveling along the feedback loop. Consequently, from this moment the dynamic flip-flop ceases the transmission of pulses to the input of the frequency divider. It is easily seen that the discussed system may be established in

the "stop" state with a 90-95% reduction of the beginning of the stop signal.

Pulse φ from the other output of core M_1 will establish all cores of the frequency divider (with the exception of the sixth core in divider D_1 and the third core in divider D_2) in the state in which pulse φ will enter 1 (see Figure 4b). Such state of the cores ensures the arrival of the control pulse from the output of frequency divider D_2 precisely at the time of reception of the middle of the pulses. Consequently, their recording will also occur at moments of reception of the middle portion. If it is necessary to record received pulses at phase-displaced moments, entry of 1 must be performed in other cores of frequency dividers D_1 and D_2. It must be kept in mind that upon displacement of entry of 1 to the left by one core in divider D_1 the selective pulses will be phase-displaced in the direction of time lag

$$T = t_0/K_1K_2,$$

where K_1 and K_2 are the frequency division coefficients of D_1 and D_2. Conversely, upon displacement of entry of 1 in cores D_1 to the right, the selective pulses will be displaced in the direction of a lag by the same magnitude. Upon displacement of entry of 1 in the cores of frequency divider D_2 the selective pulses will be phase-shifted in time

$$K_1T = t_0/K_2.$$

Thus, by varying the point of entry of 1 into the cores of the frequency divider, the selective pulses may be established in any phase relative to the received pulses. In similar manner we may vary the phase of selective pulses in the circuits shown in Figures 1 and 2 by changing the arms of the frequency divider flip-flops connected to the clearing bus with a reset pulse φ.

RECEIVER PHASING ERROR

Let us determine the phasing error of start-stop receivers operating without interruption of the control pulse generation. It must be pointed out that the circuits of start-stop receivers operating without interruption of the control pulse generator are essentially equivalent to mechanical start-stop receivers with ratchet gearing in which the frequency divider serves as the ratchet.

The relative phasing error Θ in the discussed circuits may be determined from the expression

$$\theta = \pm T/2t_0, \tag{1}$$

where T = 1/f is the period of the control pulses reaching the frequency divider input and t_0 is the elementary pulse duration.

Between the elementary pulse duration t_0 and the control pulse period T there exists the relation

$$t_0 = K_dT \tag{2}$$

where K_d is the frequency division coefficient.

In using flip-flops the frequency division coefficient is determined from the expression

$$K_d = 2^n \tag{3}$$

where n is the number of flip-flops.

Inserting t_0 and K_d from (2) and (3) into formula (1), we obtain

$$\theta = \pm \frac{1}{2K_d} = \pm \frac{1}{2^{n+1}}. \quad (4)$$

When n = 5, we obtain $\theta = \pm 1/64$ or $\theta = \pm 1.56\%$.

It is seen from formula (4) that the smaller the phasing error, the greater the frequency division coefficient K_d. However, as is seen from the example, with the use of only five flip-flops contactless receivers have a phasing error considerably smaller than mechanical start-stop receivers with friction clutches.

PROCEDURE FOR CALCULATING INTERFERENCE IN COMMUNICATION LINES FROM AC ELECTRIC RAILWAYS

M.I. Mikhaylov, L.D. Razumov and M.V. Markov

The report discusses two methods of calculating the interference of AC electrified railroads with communications circuits. The fundamentals of a simplified approximate method of calculation based on an equivalent disturbance current are presented and the areas of application of calculation on the basis of harmonic components are defined.

Interference from AC electrified railroads, as from most high-voltage lines, is due to the magnetic and electric influence of harmonic components of current and voltage of audio frequency arising in them. Harmonic components appear as the result of rectification of alternating current on electric locomotives.

Calculation and experiments show that in practically all cases of proximity of a communications line to an AC electrified railroad, the noise in the lines is caused by the influence of the magnetic field (by the magnetic influence of the harmonic components) and by asymmetry of the parameters of the communications conductors.

The resulting psophometric noise voltage in the communications circuit is equal to the square root of the sum of the squares of the voltages (between conductors of a two-wire circuit due to the influence of each harmonic component of current in the contact system) referred to a frequency of 800 cycles by multiplication by the coefficients of acoustic effect.

In the audiofrequency range the 3rd to the 35th current harmonics of the contact system are of practical importance. However, in the "Provisional Rules for Protection of Wire Communications and Wire Broadcasting

Equipment Against Interference from Contact Systems of AC Electrical Railroads" [1] it is recommended that only the 13th to 23rd harmonics be considered for a double-track railroad and the 19th to 27th harmonics for a single-track railroad.

Thus, for a double-track railroad

$$U_n = \sqrt{\sum_{\kappa=13}^{23} (p_\kappa U_{T\kappa})^2}\ \text{v} \tag{1}$$

and for a single-track railroad

$$U_n = \sqrt{\sum_{\kappa=19}^{27} (p_\kappa U_{T\kappa})^2}\ \text{v}, \tag{2}$$

where p_k is the coefficient of acoustic effect at the frequency of the k-th harmonic component as determined from the CCITT curve and U_{Tk} is the voltage between conductors of the two-wire circuit induced by the k-th harmonic component of current in the contact system.

U_{Tk} may be expressed by the following formulas in which all the quantities therein are given for the frequency of the k-th harmonic component of current of the contact system.

a) For open-wire and cable circuits of any length

$$U_{T\kappa} = 2\omega_\kappa M_\kappa I_\kappa r_\kappa \eta_{i\kappa} \frac{\operatorname{sh} \gamma_\kappa l_c \operatorname{sh} \gamma_\kappa \frac{l_e}{2}}{\gamma_\kappa \operatorname{sh} \gamma_\kappa l}, \text{v}. \tag{3}$$

b) For short open-wire lines and cables

$$U_{T\kappa} = \omega_\kappa M_\kappa I_\kappa r_\kappa \eta_{i\kappa} \frac{l_e}{2}, \text{v}. \tag{4}$$

Here $\omega_k = 2\ f_k$ (f_k is the frequency of the k-th harmonic component of current in the contact system); M_k is the mean coefficient of mutual inductance in h/km between the contact system and the communications conductor at the frequency of the k-th harmonic component; I_k is the current is amperes of the k-th harmonic component in the contact system; r_k is the resulting shielding coefficient at the frequency of the k-th harmonic component of current; for an open-wire line it represents the shielding coefficient of the rails and for a cable line the product of the shielding coefficients of the rails, the cable sheath and adjacent conductors; η_k is the noise sensitivity coefficient of a two-wire telephone circuit for the k-th harmonic component of current of the contact system; γ_k is the propagation constant per km of the communications conductor at the frequency of the k-th harmonic component; l is the total length in km of galvanically unseparated communications circuit subject to interference; l_e is the design length in km of the influence of a sector of the communications circuit. For circuits of great length l_e is taken equal to half the distance between traction substations $l_{Ts}/2$ in km; l_c is the length in km of the sector from the end of the communications circuit to the center of the sector subject to interference.

Current I_k may be determined from the formula

$$I_\kappa = I_{Ts} \kappa_f \kappa_j \kappa_m a, \tag{5}$$

where I_{Ts} is the traction substation load current required by all electric locomotives located on the examined influencing sector of the contact system, a; k_f is a coefficient characterizing the nonsinusoidality in shape

of the traction current wave; k_i is a coefficient accounting for the degree in change in interference upon a change in load current I_{Ts} of the contact system; k_m is a coefficient accounting for the degree of change in interference as a function of number (m) of electric locomotives, the load currents of which affect the comunications line.

For example, for a double-track railroad m = 5 and I_{Ts} = 1000 amp, K_I = 0.47, k_m = 0.34; similarly, for a single-track railroad, when m = 3 and I_{Ts} = 750 amp, k_I = 0.44, k_m = 0.47; k_I and k_m are independent of the number of harmonic components; k_f depends not on the load current or number of electric locomotives but on the distance between traction substations, the supply circuit of the contact system, the number of harmonic components, electrical parameters of the contact system, traction substations and electric locomotives. Figure 1 gives the values of harmonics in the contact system current of a single track and double track railroad for the above given values of I_{Ts} and m.

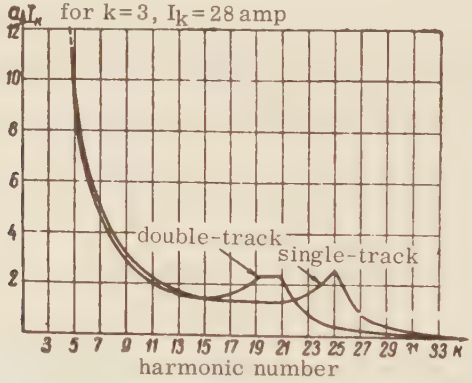

Figure 1

Calculation on the basis of harmonic components requires a knowledge of the magnitude of 5-6 current harmonics of the contact system and calculations of the noise voltage components induced by the current harmonics.

In the simpler approximate method of calculation the interfering current is taken equal to an equivalent current at a frequency of 800 cps which would have the same effect on the communications circuit (i. e., it may induce the same noise voltage) as an operating current with all harmonics. In this case, calculation of noise voltage reduces to calculation from formula (3) or (4) at only one frequency of 800 cps.

In order to use this method it is necessary to know the equivalent interfering current. The formula recommended by CCITT for determining equivalent interfering current does not give a quantitatively accurate picture of interference in two-wire telephone circuits.

As is known, the psophometric values of the interfering current is determined by measurement or by calculation from the formula

$$I'_{ps} = \sqrt{\sum_{\kappa=3}^{35} (p_\kappa I_\kappa)^2}, \text{ amp} \qquad (6)$$

Inserting into this formula the values of p_k and I_k from formula (5), we find with sufficient accuracy

$$I'_{ps} = 0{,}0365\, I_{T\pi} \kappa_l \kappa_m, \text{ amp} \quad (7)$$

Current I'_{ps}, however, may not be taken as the equivalent interfering current, since all other electrical quantities in formulas (3) and (4) — ω, γ, η, M and r — depend on frequency. In this connection it is necessary to multiply the psophometric value of the interfering current by a correction factor which takes into account the actual influence of the harmonic components of current on the telephone circuit.

This correction factor may be determined from the numerical relationship of the noise voltages calculated by both methods for different values of the quantities in formulas (3) and (4).

Thus, the ratio of noise voltage as calculated for harmonics (U_T harm) to the noise voltage calculated for psophometric current ($U_{T\ 800}$) is the sought correction factor k_n

$$\kappa_n = \frac{U_{T\,\text{harm}}}{U_{T\ 800}}. \quad (8)$$

Here ($U_{T\ 800}$) is determined from formulas (3) or (4), in which all quantities are taken at a frequency of 800 cps and, instead of I_k, I'_{ps} is taken. For example, for circuits of short length

$$U_{T\ 800} = 5000 M_{800} I'_{ps} r_{800} \eta_{800} \frac{l_e}{2}, \text{ v.} \quad (9)$$

The equivalent interfering current will be equal to the psophometric value of the interference current mutliplied by correction factor k_n

$$I_{e\kappa} = \kappa_n I'_{ps}, \text{ amp} \quad (10)$$

Below we show the possibility of using formulas (1) and (2) for a reduced number of harmonics and also present the results of calculation of correction factor k_n for a few specific conditions of influence of the single-phase 50-cycle alternating current contact system on open-wire and cable telephone circuits.

A comparison of the calculated data revealed that for open-wire lines the results of calculation with a reduced number of harmonics i.e., from formulas (1) and (2) yield values up to three times smaller than the results of calculation from the practically total number of components (from the 3rd to the 35th harmonics). For cable communications lines this departure has a maximum of 10-12% and in most cases does not exceed 5%.

Thus, for open-wire lines calculation must be performed for the total number of harmonic components and for cable lines formulas (1) and (2) may be used.

Calculation of correction factor k_n has been performed both for open-wire and for cable telephone circuits in the presence of influence of single-track and double-track electric railroads. The circuit sensitivity coefficients and the shielding coefficients of the cables as adopted in the calculation are shown in Figures 2 and 3. Calculation on the basis of harmonic components was performed with consideration of the influence of all harmonics (up to the 33rd).

Analyzing the calculated results, we may draw the following conclusions.

1. The value of the correction factor as a whole for extremely varied conditions varies but little and hardly differs from unity in one or the other direction. The correction factor varies but little with a change in conductivity

of the earth; however, in most cases there is noticed a certain decrease in k_n with an increase in conductivity.

In most cases we may also establish a tendency toward a decrease in k_n with an increase in distance between the roadbed and the communications line. This decrease is more apparent in long circuits.

2. The value of the correction factor depends on the nature of the frequency dependence of the harmonic components of the interference current.

3. The frequency characteristic of the noise-sensitivity coefficient of the circuit is of great importance. The greater the dependence of the sensitivity coefficient on frequency, the larger the correction factor.

The calculations attest to extremely complex dependence of the correction factor on various quantities. We may, however, note one general property: if any quantity entering into the formula increases with frequency, the more pronounced this dependence (that is, the more rapid the change in the discussed quantity, the greater the correction factor); if any quantity decreases with an increase in frequency, the more pronounced this dependence and the smaller the correction factor.

For example, the sensitivity coefficient increases with frequency. The more rapid this increase, the greater

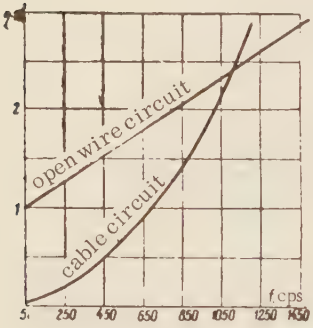

Figure 2

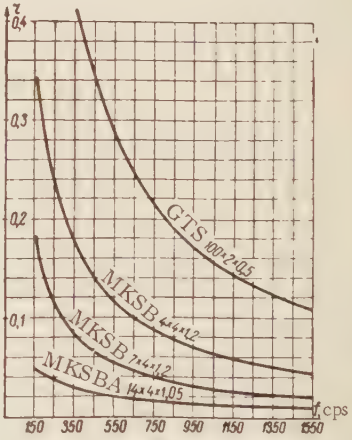

Figure 3

the correction factor. This tendency is somewhat less pronounced in cable lines and, in some cases, is even lacking. This is due to the fact that frequency dependence of the cable shielding coefficient is the reciprocal of the frequency dependence of the sensitivity coefficient of the circuit. The noted dependences of the correction factor on distance between the roadbed and the communications lines and on the conductivity of the earth are explained by the increase in the rate of decrease of the mutual induction coefficient with an increase in distance and with an increase in conductivity of the earth. Along with this, all the mentioned dependences may also be disturbed, since the collective effect of the frequency characteristics may differ.

Converting to absolute values, we may note that in most of the examined cases the correction factor varies from 0.9 to 1.2, that is, the error in calculation with the simplified formula does not exceed 20%. An

exception is the calculation for circuits of great length in which only part of the circuit is subject to the disturbance. Practical calculations are rarely performed on such circuits. For example, the table lists certain results of calculation of correction factors k_n for the influence of a double-track electric railroad on open-wire and cable communications lines.

Table 1

Conductivity of earth (mhos/m)	Circuit length (km)	Dist betw roadbed commun. circ (m)	Correction Factor				
			open-wire line	MKSB cable 4×4×1,2	MKSB cable 7×4×1,2	MKSBA cable 14×4×1,05	GST cable 100×2×0,5
1.10^{-3}	$l = l_e \leq 30$	50	1,03	1,08	1,08	—	0,96
		100	1,03	1,07	1,07	—	0,97
		500	0,98	1,02	1,02	—	0,97
		1000	0,95	0,98	0,97	—	1,02
25.10^{-3}	$l = l_e \leq 30$	50	1,01	1,05	1,05	1,08	—
		100	0,98	1,02	1,02	1,05	—
		500	0,94	0,97	0,96	0,98	—
		1000	0,92	0,94	0,93	0,94	—
25.10^{-3}	$l = 60$ $l_e = 30$	50	1,18	—	—	1,08	—
		100	1,16	—	—	1,05	—
		500	1,05	—	—	1,02	—
		1000	0,89	—	—	0,93	—
100.10^{-3}	$l = l_e \leq 30$	50	0,98	—	—	—	—
		100	0,95	—	—	—	—
		500	0,92	—	—	—	—
		1000	0,91	—	—	—	—

The performed calculations do not embrace all possible values of the quantities in the formulas. For example, it is known from numerous measurements that the sensitivity coefficients in different circuits have different values; the shielding coefficients of different cables also differ considerably.

However, the frequency dependences of sensitivity coefficients or shielding coefficients as a whole have the same character. This permits us to conclude that the derived results are applicable for other possible cases of proximity of circuits and conditions of laying communications lines.

Considering the general approximate nature of the calculations, an error of 15-20% may be considered wholly acceptable in determining noise voltage.

Then, on the basis of the tabulated data, we may assume (as is recommended in the rules" 1 , that at distances between the railroad and communications

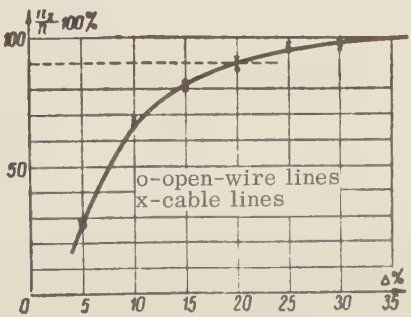

Figure 4

lines of less than 500 meters, $k_n = 1.15$ and at distances of 500 meters and greater, $k_n = 1.0$. The cumulative curves obtained with this error in calculating noise voltage from the approximate formulas are given in Figure 4.

The percentage value of error Δ is given on the x-axis, and the percentage of cases with error less than the corresponding amount on the x-axis is plotted along the y-axis. As is seen from this figure, the curves for cable and open-wire lines coincide. In 90% of the cases the error does not exceed 20%, wherein, as is seen from Figure 4, the error is in the direction of excess over the calculated value.

Psophometric calculation of the influencing current considerably facilitates and simplifies calculation of the interference on the communications line.

CONCLUSIONS

On the basis of the above observations we may draw the following conclusions:

1. As a rule, calculation of noise voltage must be performed by an approximate method based on the psophometric value of the influencing current.

2. If the value of disturbing voltage calculated by the approximate method considerably exceeds the permissible value, and considerable expenditures are required to reduce noise to the permissible value, then the disturbing voltage must be again calculated on the basis of the harmonics of the influencing current, and only from the results of repeated calculation should protective measures be adopted.

A similar work in determining k_n must be performed for power transmission lines.

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Submitted February 8, 1961

BRIEF COMMUNICATIONS

INCREASING THE POWER GAIN OF TRANSMITTER STAGES

V. A. Khatskelevich and G. V. Benben

One of the basic problems in rebuilding radio transmitters is the simplification of circuitry. This may be achieved by reducing the number of stages in the RF section of the transmitter. Thereby we achieve simplification of tuning of the transmitter, improvement of operating conditions, an increase in power indexes and sometimes, also, in the electroacoustic indexes, and facilitation of conversion to automatic control, etc.

If the tubes of the output stage (OS) of a high-power transmitter are operated at maximum efficiency, a significant result may be obtained by reducing the power of the second intermediate amplifier (IA2) and the first intermediate amplifier (IA1) (Figure 1). The stages preceding these have relatively low power and need not be considered in our discussion. Thus, the problem reduces to a maximum increase in power gain K_p of the output and intermediate amplifier stages.

The power gain of a stage, as is known, is determined by many factors: the nature of oscillations (modulated or unmodulated), the type of modulation, operating conditions, wavelength, tube type, etc. Depending on a combination of these factors, values have been established by experiment for the different cases.

Figure 1

For example, for long and medium wavelengths in a normal (common-cathode) oscillator circuit with amplification of unmodulated oscillations by a triode stage, the power gain K_p [i.e., the ratio of oscillatory power of the amplifier ($P_{\sim a}$) to its exciter power ($P_{\sim e}$)] has been recommended [1] and is now recommended [2] to be approximately 10:

$$K_p = \frac{P_{\sim a}}{P_{\sim e}} \approx 10. \quad (1)$$

The same value is recommended under the same conditions but with amplification of modulated waves (AMW). Naturally it is assumed that both these powers are given for identical modulation conditions (points), for example, telephone (T) or peak (max):

$$K_p = \frac{P_{\sim aT}}{P_{\sim eT}} = \frac{P_{\sim a}}{P_{\sim e}} \approx 10. \quad (2)$$

If the amplifier stage uses a multigrid tube (tetrode or pentode) instead of a triode, the value of K_p is taken to be 1.5-2 times greater ($K_p \approx 15$-20) and if a common-grid circuit is used K_p is assumed to be 1.5-2 times smaller ($K_p \approx 5$-7) etc., [2].

These recommendations are fundamental in preparing the block diagram and choosing the tube types for individual stages of the transmitter.

However, since they have large safety factors, they often create an incorrect impression of the actual possibilities of one or another circuit. The "latent reserves" are sometimes so great that they permit increasing the gain by almost an order of magnitude (i.e., permit obtaining a gain of 100 or greater instead of 10). There is no need to eleborate on the practical benefits to be obtained from this.

Practical workers" may often be heard declaiming against theoretical calculations of operating conditions which show the possibility of obtaining high values of K_p. They assert that these calculations are, as it were, practically unattainable. Actually, in certain cases (for example, in frequent and rapid retuning over a range of frequencies) a certain margin of exciter power is required. On the other hand, in prolonged operation at one frequency (broadcast transmitters, certain radio navigation transmitters, etc.), particularly at medium and long wavelengths, it is wholly possible to achieve in practice the results of the mentioned calculations; this is graphically illustrated by an experimental investigation performed by the authors in 1956 and discussed below.

Here grid-circuit calculations of the amplifier are of particular importance, especially for new tubes in which, as is known, the limiting factor is not the plate circuit data but the thermal conditions of the grid. In the older tube types (G-433, G-431, GK-3000) possessing negative dynamic resistance it is also desirable to have a more accurate knowledge of electrical data for the grid circuit in order to obtain maximum K_p.

For this purpose the method of calculating oscillator operating conditions described in [3] may be recommended. It is extremely simple and reliable, since it has been checked against many precise graphical calculations from actual characteristics of tube, which also serves as a guarantee of accuracy. According to this method, which, generally speaking, was developed for application to new triodes, we may also calculate the operating conditions of other tubes, including the older types. In calculating the components of grid current it is only necessary to change the empirical correction factors k_{0g} and k_{1g} [3], decreasing them in comparison with the factors obtained for new tubes by approximately 1.5-2 times, depending on the intensity of the negative-resistance phenomena, that is, to take

$$\kappa_{0g} \approx 0.35 - 0.55, \quad \kappa_{1g} \approx 0.4 - 0.6 \tag{3}$$

(the greater the negative resistance effect, the smaller these factors).

It is evident that the greatest effect in obtaining a maximum value of K_p may be obtained from tubes with small grid cirrents, e.g., the old G-433 tubes as well as certain more recent tubes (GU-4A, GU-89A, etc.), even taking into account measures adopted to eliminate the effects of secondary emission (kenotrons, bypasses, etc.). Hence it was decided to conduct an experiment to check for the possibility of achieving large values of K_p on such tubes with a high power level. A medium-wavelength multistage

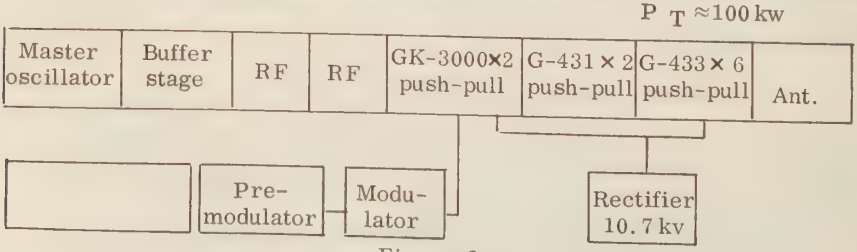

Figure 2

(seven stages) high-power AM transmitter with grid-bias modulation in the third stage from the output was used. The block diagram of this transmitter is given in Figure 2. Our discussion deals with the last three stages, which were built for push-pull operation.

Such transmitters or transmitters similar thereto (with regard to power level and power ratios in adjacent stages) were widely used in the pre-war years. With one or another modification such transmitters are still in operation at a number of stations. In particular, in the discussed transmitter changes were made in the modulator circuit, the crystal-controlled Pierce oscillator (master oscillator) was replaced with a standard VKVT-2, etc. Since these modifications are of no importance for the subject at hand, we shall not deal with them here.

The output stage (OS), operating as a modulation amplifier (MA), included six G-433 tubes which at the telephone point provided an oscillatory power of approximately $P_{\sim T} \approx 100$ kw and at the peak point (with m = 1) a power of

$$P_{\sim \max} = 4P_{\sim T} \approx 400 \text{ kw}.$$

In the second driver stage (IA2), also operating as a modulation amplifier, there were two 20-kw G-431 tubes and in the preceding (IA1) grid-modulated stage there were two 3-kw GK-3000 tubes. These stages were preceded by two RF amplifiers, a buffer stage and master oscillator, which are of no concern in this discussion.

The typical operating data for the tubes of the IA2 and IA1 stages satisfied condition (2), that is, $K_p \approx 10$. Design calculations for these stages showed a considerable reserve of power over that actually required.

On the basis of design calculations a new IA2 was built, using four 750-w GU-80 tubes (also in push-pull with $E_a \approx 2.8$ kv), which, even at reduced screen grid voltage of approximately 500 volts (instead of the rated 600 v) easily provided the previous operating conditions for the output stage. Its driver was a stage with two GU-50 tubes (push-pull) preceded by the master oscillator (Figure 3). Modulation was introduced at the control gird of IA2.

It was, of course, necessary to reject plate modulation in view of the impossibility of achieving it under continuous operating conditions of the station. The operating conditions of the discussed transmitter were the same as previously. Experimental verification of design calculations for the possibility of obtaining large values of K_p at high power levels was, for our purposes, not obligatory, although desirable. The quantitative indexes were even slightly improved, particularly the nonlinear distortion. This was evident from the shape of the modulated envelope at the output. In the original transmitter design the envelope had obvious, clearly discernible distortions; the nonlinear distortion factor reached values of

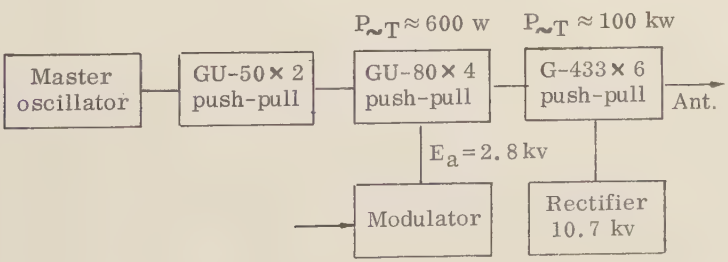

Figure 3

$K_f \approx 10\text{-}13\%$[1]. In the modified circuit these distortions were practically eliminated. Unfortunately, for technical reasons it was not possible to measure them.

The IA2 power at the telephone point was approximately 600 w, which provided an output power gain of

$$K_p = \frac{P_{\sim \text{T OS}}}{P_{\sim \text{T IA2}}} \approx \frac{100 \cdot 10^3}{600} \approx 170 \text{ (!!)}.$$

This result was obtained by the above-mentioned design calculation of the output grid circuit, particularly by calculation of its driving power as well as the power loss in the driver circuit.

In this connection there may arise the question of the convenience of listing in handbook data concerning the driving power of tubes for the purpose of using them in similar calculations. However, the numerous calculations performed by us attest to the inadvisability of this. In the first place, the driving power varies considerably with operating conditions of the stage (voltage level, cutoff angle, modulation and, particularly, the degree of utilization of the current and voltage efficiency of the tube, not to mention the various aspects of shortwave operation) and such a variety of data cannot be presented in a handbook. Moreover, existing methods of calculation are accurate and simple enough that they present no difficulties even for a technician.

Comparing the original transmitter (Figure 2) with the modified version (Figure 3), we see that the latter affords the following significant advantages:

1) Elimination of the high-power last driver stage with its large power requirements and water-cooling;

2) Elimination of the high-voltage driver before the last driver stage;

3) Increased economy;

4) Simpler circuitry;

5) Reduced transmitter dimensions (panel front);

6) Easier operation;

7) Simpler conversion to partial remote control, which is of particular interest at this time.

The above remarks on the possibility of increasing the gain of transmitter stages and of achieving a corresponding economy also apply, to varying degree, to other existing and planned transmitters. Review and reevaluation of existing ratings and circuits and realization of existing possibilities may in some cases afford significant economy of power and components.

The authors express their gratitude to engineers I. Ye. Sorokin and Ye. I. Belen'kiy for their collaboration in the experiment.

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Submitted March 1, 1961

[1] With percentage modulation m $\approx$ 90-95%.

QUANTITATIVE PREDICTION OF THE STATE OF THE IONOSPHERE

Ya. S. Rodionov and V.S. Gubenko

Short-term prediction of the state of the ionosphere is presently performed by the qualitative method [1]. Statistical extrapolation is one method by which short-term quantitative prediction may be achieved. In [2] this method tried for prediction of critical frequencies on subsequent days but did not provide encouraging results. In [3] regression equations were used for the case of a disturbed ionosphere, that is, for prediction of critical frequencies on the assumption of a disturbed ionosphere.

It our purpose below to discuss the possibility of using statistical extrapolation for prediction of critical frequencies of the F_2 layer several hours in advance. Calculations were performed for ionization density, for in this case comparison of experimental data with the results of theoretical investigation of the ionization-recombination equation does not require additional calculation.

The difficulties in statistical extrapolation lie in the inadequacy of investigations of changes in ionization density as a random process and in the fluctuational nature of this process.

For extrapolation of the ionization density we used, as the simplest operator, a linear operator expressing the anticipated value of deviation in ionization density $\Delta N_y(t_y)$ in the form of a weighted sum of observed values of deviation in preceding moments of time $\Delta N(t_k)$:

$$\Delta N_y(t_y) = \sum_{\kappa=0}^{n} A_\kappa \Delta N(t_\kappa).$$

The last observation is designated by a zero. The weighting coefficients A_k, providing the minimum mean square error in extrapolation σ_y, are determined from the system of linear equations:

$$\sum_{\kappa=0}^{n} A_\kappa R(t_\kappa, t_m) = R(t_y, t_m),$$

$$m = 0, 1, 2, \ldots, n;$$

Here $R(t_i, t_j)$ is the coefficient of correlation between the values of ionization density at moments t_i and t_j.

The ratio $\gamma = \sigma_y/\sigma$ was adopted as the criterion of prediction efficiency, where σ is the standard deviation of ionization density from the mean monthly value and σ_y is the standard deviation of the predicted values of ionization density from their actual values. If this deviation is less than unity, the chosen method of extrapolation provides good results.

The γ criterion permits a first-approximation evaluation of the required accuracy of short-term prediction under different communications conditions. In the most general case it is necessary to choose the operating frequency f_p so that its ratio to the predicted maximum usable frequency (φ) lies within definite limits, that is, the following equality must be observed

$$1 > \varphi_2 > \varphi > \varphi_1 > 0,$$

where φ_1 and φ_2 are the given limits of variation in operating frequency relative to the current MUF (maximum usable frequency).

The probability of reflection (P_0) with normal distribution for fluctuations in MUF and predicted values of MUF will be

$$P_0 = \Phi\left[\left(1 - \frac{\varphi}{\varphi_2}\right) : V_\gamma\right] - \Phi\left[\left(1 - \frac{\varphi}{\varphi_1}\right) : V_\gamma\right].$$

Here Φ is the normalized Laplace function; $V = \sigma : \overline{MUF}$ is the coefficient of variation of MUF (σ is the standard deviation of MUF and $\overline{MUF}$ is the mean value).

Possibly of greater importance than probability P_0 in Soviet communications systems is $P_{n0} = 1 - P_0$, where P_{n0} is the probability of absence of reflection. For the simplest case, in which the inequality MUF > f_p must be satisfied, we have

$$P_{n0} = \frac{1}{2} - \Phi[(1 - \varphi) : V_\gamma].$$

Let us take $1-\varphi = V$. In the absence of short-term prediction ($\gamma = 1$) this case corresponds to the optimum working frequency (OWF) and, correspondingly, $P_0 = 0.84$ and $P_{n0} = 0.16$.

Table 1 lists data which to a certain degree reflect the utility of short-term prediction and the required values of the γ criterion.

Table 1

γ	P_0	P_{n_0}	$P_{n_0}(1)$ $P_{n_0}(1)$
1,0	0,84	0,16	1,00
0,9	0,87	0,13	0,84
0,8	0,89	0,11	0,66
0,7	0,92	0,08	0,48
0,6	0,95	0,05	0,30
0,5	0,97	0,03	0,14
0,4	0,99	0,01	0,04

From Table 1 we may conclude that obtaining $\gamma = 0.7$ by means of short-term prediction permits a 50-percent reduction in P_{n0} and, consequently, it may be of practical value.

In order to take into account the fluctuating nature of the process, for each hour of the day a specific operator was calculated, that is, a separate set of coefficients A_k. All the calculations were performed for June and December, since the seasonal change in these months is weakly expressed.

The period of observation T_{n0} is assumed to be 18 hours and the period of forecast T_y six hours. The observation intervals T_0 were 1, 2, 3 and 6 hours in order to determine the dependence of prediction on this parameter. The quantity n varied accordingly from 18 to 3, while the number of equations in the system varied from 19 to 4.

The calculated results, performed on a digital computer by T. T. Kulikov, show that statistical extrapolation is useful both for June and for December of 1948, 1950 and 1953 [4] (Figures 1a, 1b and 1c). In all figures the solid lines correspond to June and the dashed lines to December.

As an index of the effectiveness of extrapolation for each month we took the quantity

$$\gamma_{av} = \frac{1}{24}\sum_{i=0}^{23}\gamma_i,$$

where γ_i corresponds to the i-th hour of the day.

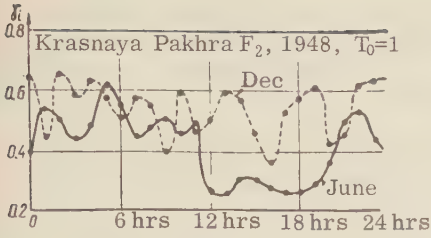

Figure 1a

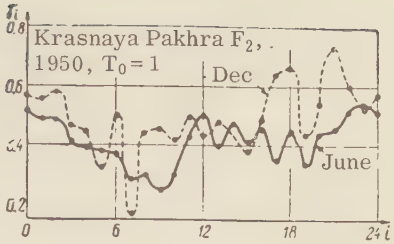

Figure 1b

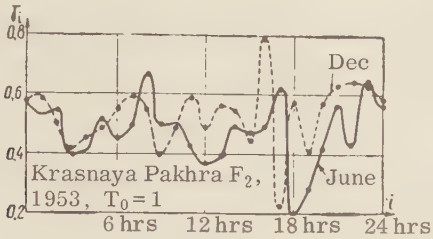

Figure 1c

Table 2

Year	Average annual No. of sunspots	γ_{av}	
		June	Dec
1948	136.2	0.42	0.54
1950	84	0.41	0.49
1953	14.3	0.47	0.53

For December the values of γ_{av} were slightly larger than for June. Dependence of γ_{av} on the level of solar activity is shown in Table 2. The results seem to indicate that γ_{av} does not have a clearly expressed dependence on the number of sunspots.

Figures 2a, 2b and 2c show the dependence of the average index of effectiveness of prediction on the period of observation. Extrapolation is most effective when $T_0 = 1$.

The calculations were used to check the possibility of linear extrapolation of ionization density, not for deviations from the monthly means but from the values of density themselves. The results show that the extrapolation error in this case is slightly greater. Figure 3 compares the two forms of extrapolation for 1950. For other years these relationships are similar. The calculations also showed that it is possible to increase the accuracy of extrapolation of ionization density if the extrapolation is performed for critical frequencies and then the predicted value of critical frequency is converted to the corresponding value of ionization density. Here, when $T_0 = 1$, for June 1950, γ_{av} was reduced from 0.41 to

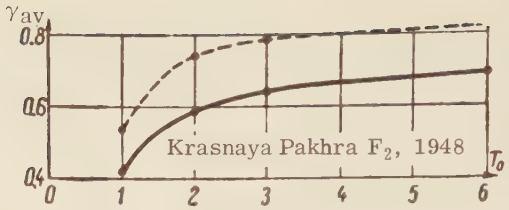

Figure 2a

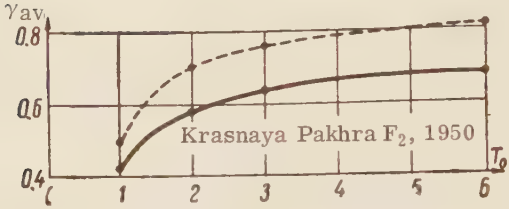

Figure 2b

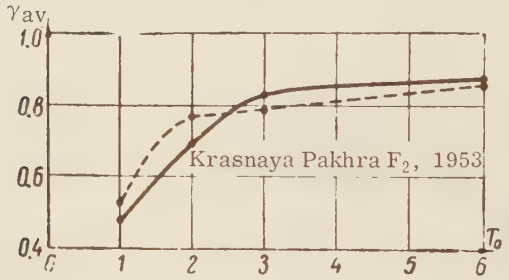

Figure 2c

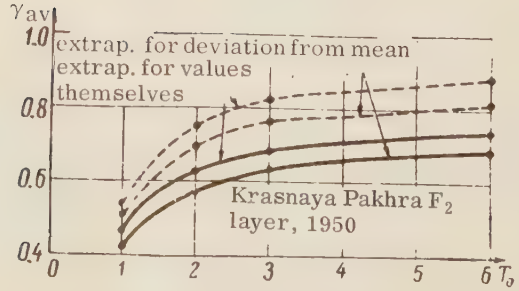

Figure 3

0.30 and for December 1950 from 0.49 to 0.47.

The results show that statistical extrapolation is deserving of attention as one of the methods of short-term prediction, although it is likely that many difficulties are still to be encountered.

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SECOND PLENARY ASSEMBLY OF THE INTERNATIONAL CONSULTATIVE COMMITTEE ON TELEGRAPHY AND TELEPHONY

The Second Plenary Assembly of CCITT was held from 8 to 16 December 1960 in New Delhi, the capital of India. Before the opening of the Assembly most of the research commissions of CCITT met to prepare the final text of recommendations subject to approval by the Plenary Assembly. The Assembly met after approximately four months of work by the Consultative Committee on Telephony and Telegraphy, organized at the First Plenary Assembly in 1956 by uniting the previously existing independent consultative committees on telegraphy and telephony.

The Second Assembly was officially opened by Jawaharlal Nehru, Prime Minister of India. Particular mention must be made of the preparations and conduct of the Plenary Assembly on the part of the government and the Department of Posts and Electrical Communications of India, which made available to the assembly a well-equipped building for the large meetings and created all the conditions necessary for efficient work by the research commissions and the assembly as well as for the recreation of the delegates.

Delegations from 54 communication administrations took part in the work of the Plenary Assembly — members of the International Communications Union (ICU), seventeen well-known private companies and eight international organizations. The total number of delegates was approximately 350. The Soviet Union was represented by delegates from the USSR, the

Ukrainian SSR and the Belorussian SSR. Approximately ten nations took part in the work of the CCITT Assembly for the first time, including such nations as Mali and Senegal, which achieved their independence only a short while ago.

MOST IMPORTANT RECOMMENDATIONS ADOPTED BY THE PLENARY ASSEMBLY

The recommendations adopted by the Plenary Assemblies of the CCITT are extremely broad in nature and provide for standardization of the chief technical characteristics of telephone and telegraph systems used for international communications, methods of operating these systems and unification of rates for the use of international communications channels.

The technical recommendations have as their purpose the standardization of various types of equipment and communications lines, methods of transmission and electrical characteristics of communications channels in the light of modern engineering achievements. The technical recommendations of CCITT are widely used by all nations in the development of equipment and construction of communications lines intended not only for international communications but also for intranational communications. The recommendations and instructions developed by the CCITT in the field of operations are intended to ensure coordination of the work of operating personnel of different countries in the organization and maintenance of international communications.

Recommendations in the Field of Transmission Systems

Recommendations in this field were prepared by the first research commission (RC) and four of its subcommissions as well as by independent operating groups on transmission of data and noise. Of the number of recommendations adopted in the report of this commission, we must note those recommendations defining the fundamental electrical characteristics of the 300-channel system of multiplexing a small-diameter coaxial cable (1.2/4.4 mm). A number of questions associated with elaboration of these recommendations remain for further study.

Recommendations for refining noise calculation procedures with conventional standard circuits 2500 km in length are of particular interest. In addition to the mean psophometric noise power of 10,000 picowatts, which must not be exceeded in the course of any given hour, the permissible mean noise per minute is defined (which must not be exceeded in 20% and 0.1% of the time in any month). Also defined are the permissible nonpsophometric (integral) noise power and distribution of noise power between individual frequency converters. A recommendation was adopted which defines the reference load of a group transmission channel as a continuous-spectrum signal which must be used in calculating mutual influence between channels of multichannel systems due to nonlinear transitions.

In the light of the increasing range of international communications lines the CCITT devoted great attention to the development of recommendations dealing with long-distance intercontinental communications. Recommendations were adopted which defined the general characteristics of international circuits longer than 2500 km (for open-wire lines up to 10,000 km and for cable lines without long submarine sections up to 25,000 km).

In accordance with the new plan for interconnection of channels, the chief criterion for stable communications is the degree of echo disturbance. On this basis the net loss of the communications channel (between four-wire terminals) in the absence of echo suppressors is established as a

function of length of communications. Moreover, for the purpose of eliminating echo disturbance, the number of series-connected four-wire circuits is limited; net loss on these circuits must be established as a function of the number of low-frequency four-wire through connections. Investigation of the question of a new plan for interconnection of communications channels on the basis of approved principles will be continued.

A number of recommendations dealing with open-wire communications systems were adopted: distribution of noise on a conventional reference circuit; definition of maximum extent of high-frequency communications over open-wire lines; standardization of pilot frequencies for three-channel systems.

In conjunction with the development in a number of countries of transmission systems employing a standard diameter coaxial cable (2.6/9.4 mm) with a bandwidth of up to 12 Mc (for 2700 telephone channels or a television channel and 1800 telephone channels) CCITT adopted a number of recommendations providing for standardization of the characteristics of this system. Investigation of individual parameters of a system, particularly with reference to simultaneous transmission along a group channel of telephone and television signals, will be continued.

A number of recommendations dealing with the transmission of broadcast programs were approved. A standard broadcast channel with bandwidth of 50-6400 cps was approved, which in the previous recommendations had been a channel of the old type which could be used in the absence of a channel with bandwidth of 50-10,000 cps. Refinements were introduced into the recommendations concerning noise and crosstalk in broadcast channels and took into account the increased range of transmission. A frequency characteristic was recommended for a preemphasis circuit in a broadcast channel and a corresponding equalizing circuit. The recommendations on permissible nonlinear distortion in broadcast channels and the table of weighting coefficients for a filter weighting network of a broadcast psophometer were revised.

A recommendation was adopted which provided for the use of terminal multiplex equipment for eight channels within the bandwidth of the primary group of 48 kc with transmission of ringing and dialing signals outside the effective transmitter bandwidth of each telephone channel. The use of equipment with spacing between virtual carrier frequencies of 6 kc must not cause limitations with regard to range of transmission. The channel characteristics must correspond to those adopted by CCITT for a system with 4 kc spacing between carriers; at a frequency of 4.3 kc the over-all channel loss must not fall below the specified value in order to avoid oscillation.

In connection with the fact that the transmission of broadcast commentaries over ordinary telephone channels is becoming increasingly common in many countries, the CCITT adopted new recommendations concerning the levels of transmission and crosstalk within channels in such cases.

Recommendations were adopted which define the conditions of use of telephone channels for facsimile transmission (types of circuit, levels, overall loss, phase distortion, noise) and range of facsimile transmission. The question of the use of vestigial-sideband transmission was left for further study.

The Plenary Assembly approved a recommendation defining the power levels required for transmission of data along leased circuits and communication circuits and also approved a recommendation concerning the equivalence of binary number notations and significant states in a two-position code.

Problems of Operation and Rates

The recommendations on these questions were prepared by the second research commission and its subcommissions. The recommendations adopted in the field of telegraph communications reflect the wide development of automatic subscriber telephone networks and those for general use.

A number of recommendations dealt with the use of reperforators in automatic telerepeaters. The mean efficiency of connections in an automatic general-purpose system (Gentex) was defined.

Revisions were made on the regulations for servicing of automatic subscriber telegraph (Telex) and the order of mutual calculations between countries in the use of subscriber telegraph networks was defined. A recommendation was adopted for changing Alphabet No. 2 for the purpose of freeing two combinations for internal use by individual countries.

Concerning problems of operation and tariffs in international telephone communication, a number of recommendations were adopted for apportionment of charges in automatic transit, on establishing charges for wrong-number connections, on not permitting access of subscribers of one country to the information operators of another country, and a number of other problems.

Radio Relay Links and Radio Communications by Wire

The Plenary Assembly approved a number of recommendations prepared by the third research commission defining the sequence of interconnection of systems using different methods of signalling, connection of cable systems and radio relay systems along tertiary and secondary groups and television channels of radio relay and coaxial cable links.

Technical Operation

The Assembly approved the considerably revised regulations and instructions prepared the fourth research commission on technical operation of international telephone, radio television broadcast channels.

Protection of Communications Lines

The first three sections of the directive on protection of lines against electromagnetic disturbances, prepared by the fifth research commission, were approved as were a number of sections of new recommendations for protection of underground cables against corrosion, as prepared by the sixth research commission.

Definitions, Dictionary and Symbols

With regard to the report by the sixth research commission the Plenary Assembly adopted a resolution to publish in the near future a revised list of basic definitions of electrical engineering in English and French and to begin work of preparation of a new, completely revised and expanded list of basic definitions which will be published in French, English, Russian and Spanish.

Telegraph and Facsimile Equipment

On the basis of proposals by the eighth research commission the recommendations adopted by CCITT establish norms for the operation of start-stop teleprinters and norms for the receiving section of synchronous systems and for synchronous systems with automatic error correction. Recommendations were adopted defining specifications for regenerative relays and transponders. Changes were made in the recommendations standardizing the basic parameters of facsimile equipment and a recommendation was adopted for a test pattern for facsimile communications.

Quality of Telegraph Transmission and Telegraph Channels

A number of recommendations prepared by the ninth research commission were adopted for standardization of frequency telegraph systems and procedures for evaluating qualitative indexes of these systems.

In particular, recommendations were adopted defining a method of separate measurement of random and characteristic distortions and of bias distortion within a channel and a procedure for preventive measurement of telegraph channel distortion. A formula was recommended for adding distortion in series connection of channels. Recommendations were adopted for standardization of voice-frequency telegraph systems with frequency modulation, with amplitude or frequency modulation with carrier spacing of 240 cps and infraacoustic telegraph systems composited with a three-channel high-frequency telephone system. Telegraph signal distortion norms were given for measurements made with a start-stop distortion meter.

Switching of Telegraph Channels

These recommendations were prepared by the tenth research commission. Among the most important recommendations were: signals in "Telex" international service, standardization of "Telex" signals in radio channel operation, restrictions against deliberate improper connections in automatic "Telex", standardization of "Gentex" signals, and others.

Semiautomatic and Automatic Long-Distance Telephone Communications

Among the recommendations approved by the Plenary Assembly on the basis of the report of the eleventh research commission on problems of automation of long-distance telephone communications, the following must be mentioned: measures for reduction of noise at telephone exchanges; conditions for interconnection of three automatic international circuits; characteristics of signaling systems outside the speech band for multiplexing equipment with eight-channel primary groups; permissible bandwidth of voice-frequency signals of existing automatic telephone exchanges; on the number of digits in a national numbering system in non-European continents; measures which must be adopted in automatic service during prolonged overloading at a through station.

NEW STRUCTURE OF CCITT

The CCITT structure existing in the period between the first and second plenary assemblies was determined under conditions of a lack of experience in operation of a unified telegraph-telephone committee. Hence, after the first four years of operation of such a committee, and also taking into account modern trends in development of communications techniques, it proved necessary to introduce substantial changes in the organization of CCITT. The organizational structure of CCITT as adopted by the Second Plenary Assembly, with an indication of the new and corresponding old research commissions, is given in the table. A distinctive feature of the new structure is the absence of subcommissions within the technical commissions, which made the CCITT organization unwieldy and was an impediment to the development of recommendations.

One new aspect of the CCITT organization is the creation of three special research commissions (A, B, C) which study problems of interest to a number of other research commissions. The creation of these commissions (on problems of data transmission for computing centers, on problems of investigation of noise in transmission systems and on problems of a worldwide system of semiautomatic and automatic telephone communications) reflects the developmental trends in communications engineering.

TECHNICAL ASSISTANCE TO NEW AND EMERGENT NATIONS

Much of the work of the plenary assembly was devoted to the question of the organization and types of activity of CCITT in giving technical assistance to new and emergent nations. The necessity for expanding this assistance was indicated in a special recommendation adopted in 1959 by the Plenipotentiary Conference of the International Electrical Communication Union. The CCITT assistance to new and emergent nations must be achieved on the basis of existing recommendations of CCITT and by investigation of new problems of specific interest to the mentioned countries.

The assembly decided that technical assistance to new and emergent nations will be achieved through a commission on the plan for development of an electrical communications network and its regional subcommissions (for Asia, Africa and Latin America). The regional subcommissions, in addition to development of recommendations for expansion of electrical communications networks in appropriate regions of the world, will define the technical, operational and rate-establishing problems of interest to the countries of one or another region. In addition, if for a given problem CCITT has already made recommendations, the subcommission must present them in a form suitable for use by the appropriate nations. In the absence of developed recommendations a problem may be referred to one of the existing technical research commissions of CCITT or CCIR, which may create a special working group for this purpose.

It is also provided that new and emergent nations be given technical literature.

WORK PROGRAM OF RESEARCH COMMISSIONS IN 1961-1964

The Plenary Assembly approved a list of problems to be studied by research commissions of CCITT in the period before the next assembly. During the next few years the focus of CCITT's attention will be directed to investigation of the following problems.

In the Field of Transmission Systems

Characteristics of a 300-channel system for multiplexing of a small-diameter coaxial cable; definition of the characteristics of a "standard" system for multiplexing a symmetrical cable with transistors; frequency and phase characteristics of telephone channels of extreme length (for worldwide communications); use of communications channels for transmission of digital information (data); methods to be recommended for ensuring standards for noise and crosstalk in broadcast channels with distance of transmission up to 2500 km; evaluation of nonlinear distortion of compandors in simultaneous application of two frequencies (with consideration of the use of compandors in telephone channels with a two-frequency system of signaling); characteristics of a system for multiplexing of coaxial cable with bandwidth up to 12 Mc to be used for simultaneous transmission of telephone and television signals; an eight-channel system for multiplexing of open-wire lines; critical values for over-all attenuation of a telephone channel outside the effective bandwidth of 300-3400 cps; measurement of nonlinear distortion in multichannel systems by means of a continuous-spectrum signal; electrical characteristics of compandors for telephone channels; telephone communications systems with pulse-code modulation; a new plan for interconnection of circuits from the point of view of transmission (for Europe and for a worldwide network).

In the Field of Telegraph Equipment, Channels and Switching

ORGANIZATIONAL STRUCTURE OF CCITT ADOPTED BY THE SECOND PLENARY ASSEMBLY RESEARCH COMMISSIONS

New Commission No.	Previous Commission or Subcommission	Name of Commission	Chairman	Vice Chairman
I	2/1	Telegraph operation and tariffs	Perri (Holland)	Varg (France)
II	2/2	Telephone operation and tariffs	Terras (France)	Bolchandani (India)
III	2	General principles of tariff schedules	Langenberger (Switzerland)	Garrido (Spain)
IV	4	Technical operation of international communications network	Valloton (Switzerland)	Postel'niku (Rumania)
V	5	Protection against electromagnetic disturbances	Ridel (FRG)	Mikhaylov (USSR)
IV	6	Protection of cable sheath and open-wire supporting structures	Holstrom (Denmark)	Muktadir (Pakistan)
VII	7	Definitions, dictionary and symbols	Gella (Spain)	Bidzhi (Italy)
VIII	8	Printing-telegraph equipment	Kerr (Australia)	Savitskiy (Ukrainian SSR)
IX	9	Quality of telegraph transmission and telegraph channels	Roque (France)	Renton (Britain)
X	10	Telegraph switching	Jansen (Holland)	Fogera (France)
XI	11	Telephone switching and signaling	Tobin (Britain)	Vasil'yev (USSR)
XII	12	Quality of telephone transmission and local telephone systems	Svedenborg (Sweden)	Kroutl (ChSSR)
XIII	2/4	Semiautomatic and automatic telephone systems	Lambiot (Belgium)	Chauve (France)
XIV	8	Equipment and channels for facsimile telegraphy	Fialkovskiy (Poland)	Bitter (FRG)
XV	1/1	Transmission systems	Job (France)	Galiardi (Italy)
XVI	1/2	Construction of international telephone networks	Franklin (Britain)	Kleys (Belgium)

New Commission No.	Previous Commission or Subcommission	Name of Commission	Chairman	Vice Chairman
Spec. A	Working group 43	Data transmission	Rhodes (Britain)	Vogan (US)
Spec. B		Worldwide semiautomatic and automatic telephone network	Bloker (US)	Burel (Sweden)
Spec. C	Working group "noise"	Noise (Combined CCITT and CCIR)	Williams (Britain)	Chairman of CCIR
CMTT	CMTT	Long-distance television transmission (combined CCITT and CCIR)	Angle (chairman of CCIR)	Franklin (Britain)
		Chief commission	Antinori (Italy)	Hamid (Pakistan)*
		Subcommission on plan for Asia	Vasudevan (India)	Matsuda (Japan)
		Subcommision on plan for Africa	Tedros (Ethiopia)	Mili (Tunis)
		Subcommission on plan for South America	Nunez (Mexico)	Ospina (Colombia)
		Temporary working group on intra-American telephone network	Nunez (Mexico)	Teyeda (Venezuela)

*Subject to approval by CCIR.

Remote measurement of the margin of telegraph equipment; a new telegraph alphabet; combined operation of synchronous systems; standardization of telegraph equipment with signaling speeds greater than 50 bauds; a change in recommendation F-60 on standardization of transponders; inherent distortion of elements of the telegraph channel; determining distribution of individual isochronous and start-stop distortions; investigation of random distortions in channels; distortion with instantaneous changes in level; standardization of telegraph channels for speeds in excess of 50 bauds; investigation of additive effects of distortion for channels with frequency modulation; characteristics of voice-frequency telegraph channels for operation via radio channels; introduction of the concept of operational stability of channels; operational measurement of voice-frequency telegraph channels; standardization of signals in the "Telex" and "Gentex" services and in communication through synchronous systems with automatic error correction (continuation of study); switch gear for special communications facilities (transmission of data, high-speed transmission).

In the Field of Facsimile Telegraphy

The phase correctors for facsimile channels; the use of frequency modulation for facsimile; the use of single-sideband transmission; evaluation of the quality of facsimile transmission by the use of a test pattern.

In the Field of Semiautomatic and Automatic Telephone Communication

Systematization of automatic telephone networks (a system for construction of a network, direction of traffic, etc.); changes in operating regulations and conditions taking into account expansion of semiautomatic and automatic communications; preparation of a world wide numbering system; devices required in signaling and switching equipment to permit connection and disconnection of echo suppressors when operating conditions require.

In the Field of Technical Operation

Stability of over-all attenuation of telephone channels; brief interruption of service; automatic gain control in primary and secondary groups of channels; automatic measuring devices for operational testing of automatic communications; technical operation of television channels; equipment for operation of broadcast channels; specifications for technical operation of systems operating with solid-state devices; mass periodic testing of channels; grounding of equipment.

The work program of CCITT also included a number of other problems.

CONCERNING THE THIRD PLENARY ASSEMBLY OF CCITT

The next plenary assembly of CCITT is to be held in 1964. Two proposals were made for the site of this assembly: the USSR proposed that the Third Plenary Assembly be held in Moscow and the Colombian delegation proposed that it be held in Bogota. A tally of secret ballots showed that the delegations of 25 countries voted to hold the next plenary assembly in Moscow and 18 countries voted that it be held in Bogota. Thus, the Third Plenary Assembly will be held in the capital of the USSR, Moscow.

AUTHOR'S CERTIFICATES

Class $21a^1$, 11_{01}, No. 139683. Related to Author's Certificate No. 136764. B.P. Terent'yev. Electronic teleprinter.

For the purpose of printing telegrams on strip paper with roll feed it is proposed to use a moveable shaft-mounted type drum with characters and strikers actuated by electromagnets.

Class $21a^1$, 32_{11}, No. 140085. V.G. Patrunov and M.G. Artyunov. Method of obtaining magnetographic images.

Class $21a^1$, 32_{35}, No. 140086. I.I. Ilisayevskaya. Television camera tube.

Class $21a^1$, 32_{41}, No. 140087. V.A. Lutskiy and V.A. Sidorko. Device for readout of a photodiode.

For the purpose of increasing operational reliability of the device it is proposed that the photodiode cell be made in the form of a voltage divider with an active resistance and the resistance of the photodiode, in parallel with which there is connected an RC differentiator, to the resistor of which the load is connected via a diode.

Class $21a^1$, 32_{54}, No. 140088. A.Z. Kan. Device for transfer of the image of a scanning spot from a luminescent screen.

Class $21a^1$, 34_{11}, No. 140092. Yu. Ye. Karpeshko. Method for correction of half-tone characteristics in facsimile equipment.

For the purpose of improving image quality it is proposed that correction be used only for signals of large image details, for which purpose these signals are separated from fine signals details by upper and lower frequency filters, whereupon the low-frequency signals are corrected and the high-frequency signals pass through a variable-gain amplifier and then these and the other signals are combined.

Class $21a^1$, 34_{11}, No. 140093. S.I. Kochergin. Device for simultaneous transmission of image elements.

For the purpose of simplifying and increasing operating reliability of the device it is proposed that it be made of a layer of a ferrodielectric, one of the plates of which is a mosaic with a number of cells equal to the number of image elements per line and the other plate of which is a continuous conducting layer connected to the load resistance while the mosaic cells are connected to sine-wave voltage generators of different frequency.

Class $21a^1$, 34_{22}, No. 139684. G.V. Braude. Method of nonlinear aperture compensation.

For the purpose of decreasing the noise level during aperture compensation it is proposed that the amplitude characteristic of the channel for the main signal of the aperture compensator be made nonlinear so that the slope of the amplitude characteristic is decreased in transition from sections corresponding to the dark points of the image to the bright points and that the resulting nonlinear distortion be corrected by a gamma correction unit.

Class $21a^1$, 34_{22}, No. 140094. A.A. Gol'din and T.V. Shmakov. Electronic method of printing combined frames.

For the purpose of improving the quality and simplifying the printing process it is proposed that there be used a cathode-ray tube which creates

a light flux with two different spectral components, one of which is used for analysis of the image to be printed onto another film and the other for printing of this image.

Class $21a^1$, 36. No. 140095. S.M. Persin. Device for nonlinear pulse-time coding achieving functional dependences on one or two variables.

For the purpose of simplifying the device it is proposed that linear compensating networks be used with transient characteristics corresponding to the required static nonlinear dependences in combination with a comparator serving to measure the output pulse determined by the interval between the beginning of comparison and the moment of coincidence of the input signal and the signal from the transient characteristic of the compensating network.

Class $21a^1$, 36. No. 140096. I.M. Zhovtis. Relaxation oscillator.

Class $21a^1$, 18_{01}, No. 140098. V.D. Chelyshev. Method of regulating cathode currents of two or more radio tubes.

In regulation by means of a connection between the stabilizing resistor and the tube cathodes and connection of the control grid of one tube with the cathode of another for the purpose of isolating the signal circuits and the cathode current regulating circuit with simultaneous application of self-bias to the control grid of any tube it is proposed that a stabilizing resistor be connected to the cathode of each tube in series with the self-bias resistor and that the control grid of one tube be connected with the stabilizing resistor in the cathode circuit of the other tube through a leak resistor.

Class $21a^4$, 6_{02}, No. 14099. I.N. Prutnikov and M. Ye. Mazurov. Regenerative frequency divider.

Class $21a^4$, 8_{02}, No. 140100. G.B. Al'tshuler, A.D. Vaytsekhovskiy and V.A. Prokhorov. Quartz oscillator with contactless switching of two frequencies.

Class $21a^4$, 27, No. 140101. A.I. Ioffe and Ye. P. Cherkasov. Device for converting a continuous change in voltage phase into a continuous-current relay signal.

Class $21a^4$, 29_{01}, No. 140102. V. Yu. Tolubeyev. Single-frequency receiver for voice-frequency dialing and ringing.

For the purpose of increasing the protection of a transistorized receiver against speech currents it is recommended that it include a selective limiter tuned to the ringing-current frequency. This limiter is a series resonant circuit in the collector circuit of an amplifier transistor and connected with two back-to-back diodes joined with a resistor divider.

Class $21a^4$, 35_{14}, No. 139686. S.I. Bugrov and A.I. Pazyuk. Device for voltage or current regulation.

Class $21a^4$, 64_{05}, No. 140103. O.N. Tereshin and A. Ye. Sokolov. Method of suppressing surface currents.

Class $21a^4$, 71, No. 139689. M.S. Yanovskiy and B.N. Knyaz'kov. Polarization frequency converter.

For the simultaneous derivation of two sidebands with adjustable amplitude ratio it is proposed that the converter consist of four series-connected waveguide sections. The extreme sections have tapered transitions from a rectangular waveguide to a circular waveguide with absorption plates parallel to the broad dimension of the waveguide. The two middle waveguide sections are circular with dielectric plates, one of which is uniformly rotated.

Class $21a^4$, 71, No. 139690. V.F. Vzyatishev. Method of measuring the dielectric constant and power factor angle of high-grade dielectrics.

Class $21a^4$, 71, No. 140104. A.I. Brodskiy. Film bolometer

(Thermistor) for power measurement in the millimeter range.

Class $21a^4$, 72_{01}, No. 140106. L. Ya. Nagornyy. Transistorized AC switch.

For the purpose of increasing stability of the signal transfer constant it is proposed that the high-inductance primary winding of a transformer be connected to the emitter circuit of a transistor and the collector of this transistor be connected to the emitter of a second transistor, the base of which is connected to the negative of the power supply through a voltage divider.

Class 21b, 17, No. 139691. V.N. Novoderzhkin and V.I. Szhina-Kolobova. Method of increasing stability of operation of negative plates of a lead storage battery.

Class 21c, 63_{03}, No. 139704. G.B. Bagdanov. Voltage regulator.

Class $21d^2$, 12_{03}, No. 139719. A.V. Donskoy and G.V. Ivenskiy. Thyratron frequency divider.

For the purpose of simplifying the circuit and increasing reliability of a thyratron frequency divider for alternating voltage it is proposed that the device consist of two thyratrons (with load resistors) connected to one winding of a power transformer. The thyratron grid circuits contain storage elements consisting of resistors and capacitors in parallel.

Class $21d^2$, 12_{03}, No. 139720. B.I. Rabinovich and A.V. Shcherbak. Device for protecting transistorized self-excited inverter.

In order to increase the operating speed and reliability of a device for protecting a transistorized self-excited inverter it is proposed that it include a flip-flop by means of which a driver oscillator is cut off under emergency conditions.

Class $21d^2$, 12_{03}, No. 139721. O.G. Bulatov, V.V. Siritsa and V.P. Shipillo. Device for grid control of ionic rectifiers.

Class $21d^2$, 14_{03}, No. 139723. I.I. Kanter. Device for control of ignitron frequency converter without control grids.

Class $21d^2$, 53_{02}, No. 139731. A.N. Matyush. Parametric current regulator.

Class 21e, 11_{12}, No. 139734. D.G. Kats. Method of determining nonlinearity of beam scan on screen of cathode-ray tube.

Class 21e, 36_{03}, No. 139739. V.V. Sokolovskiy and V.F. Romanov. Digital phase meter for measurement of phase shift.

Class 21e, 36_{03}, No. 139740. M.A. Rakov. Device for obtaining 90° phase shift between two alternating voltages.

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BRITISH PATENT, Class 40(7), No. 853600, 9.11.60. Broadband directional microwave antenna for 100-250 Mc (Wideband 3.3:1). Andrew Alford. Improvements in antenna systems.

USA PATENT, Class 340-347, No. 2922151, 19.01.60. Circuit for conversion of AM signal to a binary pulse-code-modulated signal. Paul A. Reiling (Bell Telephone Labs, Inc.). Translating circuits.

USA PATENT, Class 30-65, No. 2944221, 5.07.60. Output stage of small high-power transmitter for mobile communications systems. Gilbert G. Robinson and James H. Madden. Transmitter final with series-tuned, self-neutralized tank circuit.

USA PATENT, Class 250-20, No. 2947860, 2.08.60. Emanuel P. Alvernez (Jennings Radio Manufacturing Corp.). Sideband selection. Proposes a circuit for separating either sideband with only one bandpass filter, having a passband corresponding to the bandwidth of the separated sideband.

USA PATENT, Class 250-13, No. 2948808, 9.08.60. Karl L. Neumann, and Charles E. Schneider (Radio Corp. of America). Single-sideband communications system. Proposes an improved system for single-sideband communications with a new AG circuit ensuring noise limiting and noise-free gain control. Describes a transceiver for mobile single-sideband communication at 3-15 Mc.

SWISS PATENT, Class $21a^4$, 29/01, No. 171228, 3.05.60. Detector for high-frequency signals consisting of carrier frequency and one sideband. S.O. Engstrom (Telefon AB L.M. Ericson).

USA PATENT, Class 250-20, No. 2947861, 2.08.60. Meredith S. Ulstad, and Howard D. Hern (Collins Radio Co.). Diversity combiner control system. Proposes a combining circuit in diversity reception of FM signals by a receiver having automatic passband control. The circuit establishes different threshold levels for different signal amplitudes.

USA PATENT, Class 250-20, No. 2951152, 30.08.60. Automatic phase control system for reception of AM or FM signals in the process of adding them in space-diversity or frequency-diversity reception. William Sichak and Robert T. Adams (Internation Telephone and Telegraph Corp.). Radio diversity receiving system.

USA PATENT, Class 250-20, No. 2949533, 16.08.60. Emanuel F. Read (Collins Radio Co.) Automatic gain control circuit for use in transistor amplifiers. Proposes a transistor receiver circuit ensuring constant output voltage with a change in input voltage over wide range.

USA PATENT, Class 250-212, No. 2951163, 30.08.60. Reflectors directing light rays to the solar battery of a radio receiver. Marlin R.

Shaffer and Robert L. Weisbrich (Hoffman Electronics Corp.). Versatile solar-powered radio receiver.

FRENCH PATENT, Class H 03 f, No. 1238790, 11.07.60. Parametric amplifier. (Cie Generale de Telegraphie Sans Fil). Proposes a design of parameteric amplifier for 1250-1350 Mc with booster frequency of 2500-2700 Mc with capacitive diodes and strip waveguides.

GERMAN (FRG) PATENT, Class $21a^4$, 49, No. 1050842, 13.08.59. Device for removal and insertion of channels in multichannel time-modulated systems. Helmut Oberbeck (Telefunken G. m. b. H.).

USA PATENT, Class 178-43, 5, No. 2946851, 26.07.60. Ernest R. Kretzmer (Bell Telephone Laboratories, Inc.). Television system having reduced transmission bandwidth. Proposes separation of the system into two parallel narrowband channels: (1) with a large number of steps (128) for parts of the image without fine detail and sharp drops in brilliance; (2) with a small number of steps (4) for fine details. The electron beam switches channels according to the type of image.

USA PATENT, Class 178-7, 2, No. 2955159, 4.10.60. Facsimile transmission based on television principles. Claude E. Jones, Jr. (International Telephone and Telegraph Corp.). Narrowband video communication system.

FRENCH PATENT, Class H 04 d, H 04 n, No. 1233431, 12.10.60. Community television antenna. C.C. J. Rochu. In the proposed circuit for a common antenna (which should ensure a saving of coaxial cable) all television receivers are connected to the antenna lead-in by the shortest path. The junction box contains a 2 $\mu\mu$f capacitor in series with a stepdown transformer which is tunable by means of a magnetic or brass core, permitting emphasis of video or audio signals.

BRITISH PATENT, Class 40(6), No. 856045, 14.12.60. Circuit for increasing noise immunity of vertical sync in television receiver. Paul F. Coff (The Plessey Co., Ltd.). Improvements in or relating to circuits for triggering synchronized oscillators.

JAPANESE PATENT, Class 96 D 3, No. 666, 2.02.60. Method of decreasing crosstalk in the low-frequency channel of a system with time-division of channels. (Otuau Tocuo Khitati seysakudze). Describes a method based on series phase shift of an 8-kc current with subsequent biasing with the low-frequency current of each channel.

GERMAN (FRG) PATENT, Class $21a^2$, 41/07, No. 1074089, 28.07.60. Device for testing equipment of long-distance communications systems with remotely powered repeaters. Andre Ponthus, and Georges Comte (Cie Generale d'Electricite).

USA PATENT, Class 333-28, No. 2933702, 19.04.60. Bruce P. Bogert (Bell Telephone Labs, Inc.). Time reversal delay distortion corrector. For automatic correction of time distortion introduced by communications lines it is proposed that at the center and at the end of the line there be connected synchronous devices achieving inversion during transmission of signals.

BRITISH PATENT, Class 40(4), No. 853033, 21.11.60. System for remote powering of unattended transistorized repeaters. Brian Stanley Helliwell (British Telecommunications Research Ltd.). Improvements in power feeding arrangements for telecommunications systems.

FRENCH PATENT, Class H 03 f, No. 1231211, 27.09.60. Amplifier with AGC for high-frequency communications. Proposes an amplifier with AGC to be achieved by means of a pilot frequency. It is shown that the control limits of the AGC are several nepers with an output voltage change of less than 10%.

GERMAN (GDR) PATENT, Class $21a^2$, 39/10, $21a^4$, 14/01, No. 20051, 10.10.60. Single-sideband modulator for conversion of relatively wide and relatively narrow bandwidths. Robert Dallemagne. Proposes a single-sideband modulator in which one sideband is separated by means of a phasing circuit. The distinctive feature of the modulator is the possibility of simultaneous conversion of a wide speech bandwidth and a narrow ringing-signal bandwidth without mutual interference.

BRITISH PATENT, Class 40(4), No. 856130, 14.12.60. System for automatic identification of number of automatic telephone subscriber. John Henry Keating, Eric Bertram Ireland and Leslie James Brownell (Automatic Telephone and Electric Co., Ltd.). Improvements in or relating to signalling systems.

JAPANESE PATENT, Class 96 E 32, No. 9603, 29.10.59. Circuit for registering long-distance calls. Nogami Kunisige and others (Khitati Seysakudze).

GERMAN (FRG) PATENT, Class $21a^1$, 12, No. 1074626, 28.07.60. Circuit for alternate operation of two or more transmitters over one telegraph line. Walter Herwig (Standard Elektrik Lorenz Akt.-Ges.).

USA PATENT, Class 178-7, 1, No. 2919304, 29,12.59. Charles Jelinek, and John H. Hackenberg (The Western Union Telegraph Co.). Facsimile transmitting system. Proposes a contrast circuit for facsimile transmission of graphic textual material which permits the operator to select the transmission condition at which inadequate contrast of the transmitted original is corrected and to ensure the possibility of transmission from blueprints and negatives.

USA PATENT, Class 179-100, 1, No. 2942070, 21.06.60. Laurens Hammond and others (Hammond Organ Co.). Means for binaural hearing. The device is intended to improve the apparent quality of music during playback from a single-channel recording or radio reception and should create the impression that the sound arrives uniformly from all directions.

USA PATENT, Class 179-1, No. 2948778, 9.08.60. Device for lowering the resonant frequency of a dynamic loudspeaker and decreasing its amplitude distortion. Watner W. Clements. Sound reproducing means.

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